

US007466191B2

(12) **United States Patent**  
**Kang et al.**

(10) **Patent No.:** **US 7,466,191 B2**  
(45) **Date of Patent:** **Dec. 16, 2008**

(54) **SELF REVERSE BIAS LOW-POWER HIGH-PERFORMANCE STORAGE CIRCUITRY AND RELATED METHODS**

5,856,951 A \* 1/1999 Arimoto et al. .... 365/226  
7,046,576 B2 \* 5/2006 Kim et al. .... 365/230.05

(75) Inventors: **Sung-Mo Kang**, Santa Cruz, CA (US);  
**Seung-Moon Yoo**, Austin, TX (US)

\* cited by examiner

(73) Assignee: **The Regents of the University of California**, Oakland, CA (US)

*Primary Examiner*—Kenneth B. Wells

(\*) Notice: Subject to any disclaimer, the term of this patent is extended or adjusted under 35 U.S.C. 154(b) by 39 days.

(74) *Attorney, Agent, or Firm*—Stephen C. Durant; Duane Morris LLP

(57) **ABSTRACT**

(21) Appl. No.: **11/286,197**

An integrated circuit is provided comprising a first NMOS transistor; a first PMOS transistor; a second NMOS transistor; a second PMOS transistor; a first bias voltage node coupled to a first source/drain of the first NMOS transistor; a second bias voltage node coupled to a first source/drain of the second PMOS; a third bias voltage node coupled to a gate of the first PMOS transistor; a fourth bias voltage node coupled to a gate of the second NMOS transistor; a pull-up node coupling a second source/drain of the first NMOS transistor to a first source/drain of the first PMOS transistor; a pull-down node coupling a second source/drain of the second PMOS transistor to a first source/drain of the second NMOS transistor; an input node; a storage node coupling a second source/drain of the first PMOS transistor to a second source/drain of the second NMOS transistor; an output node; an input switch coupled to controllably communicate an input data value from the input node to a gate of the first NMOS transistor and to a gate of the second PMOS transistor; and an output switch coupled to controllably communicate a stored data value from the storage node to the output node.

(22) Filed: **Nov. 22, 2005**

(65) **Prior Publication Data**

US 2006/0083052 A1 Apr. 20, 2006

**Related U.S. Application Data**

(62) Division of application No. 10/402,472, filed on Mar. 27, 2003, now Pat. No. 6,992,915.

(60) Provisional application No. 60/368,392, filed on Mar. 27, 2002.

(51) **Int. Cl.**  
**G11C 5/14** (2006.01)

(52) **U.S. Cl.** ..... 327/544; 327/530

(58) **Field of Classification Search** ..... 327/530,  
327/544

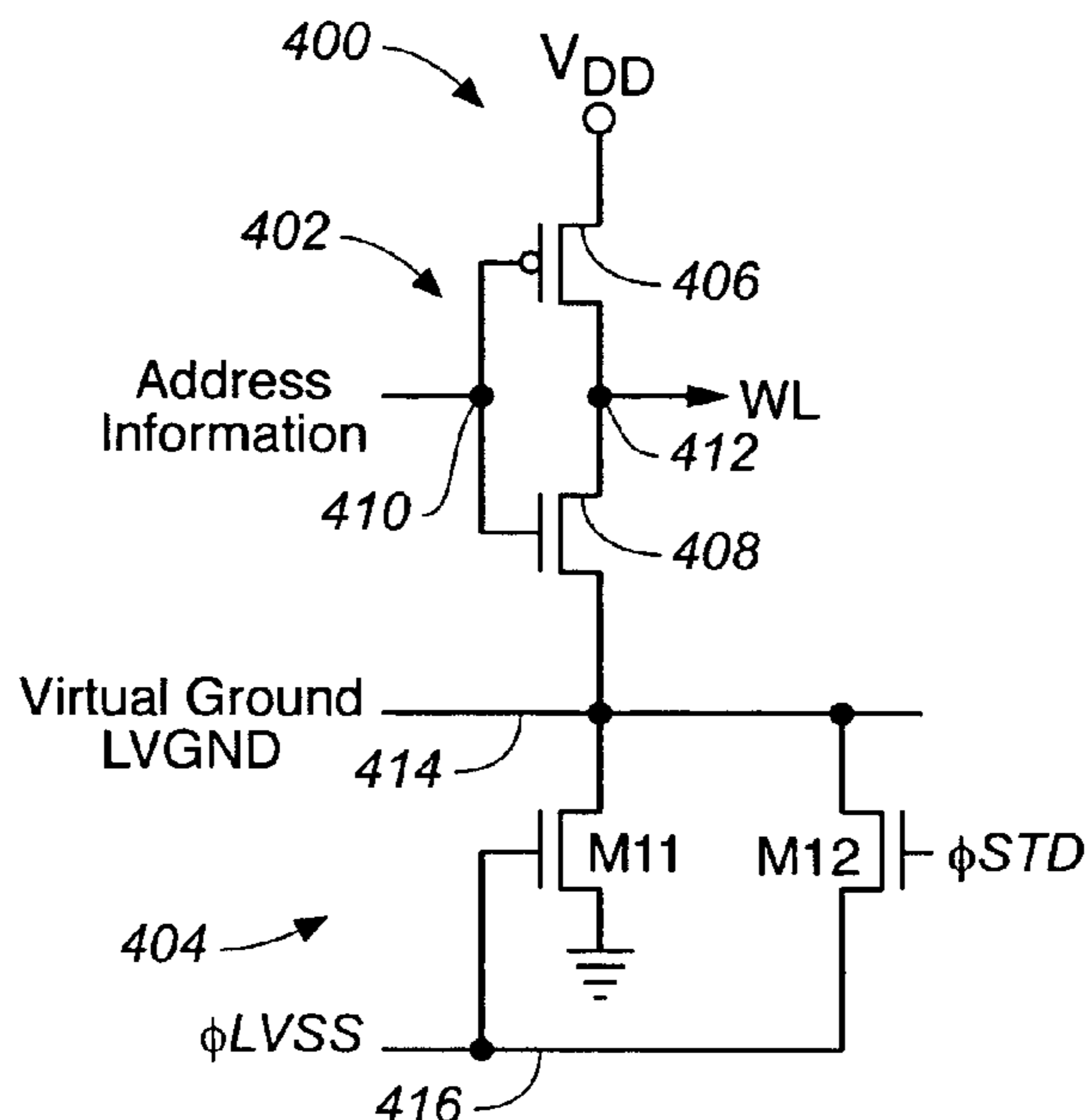
See application file for complete search history.

(56) **References Cited**

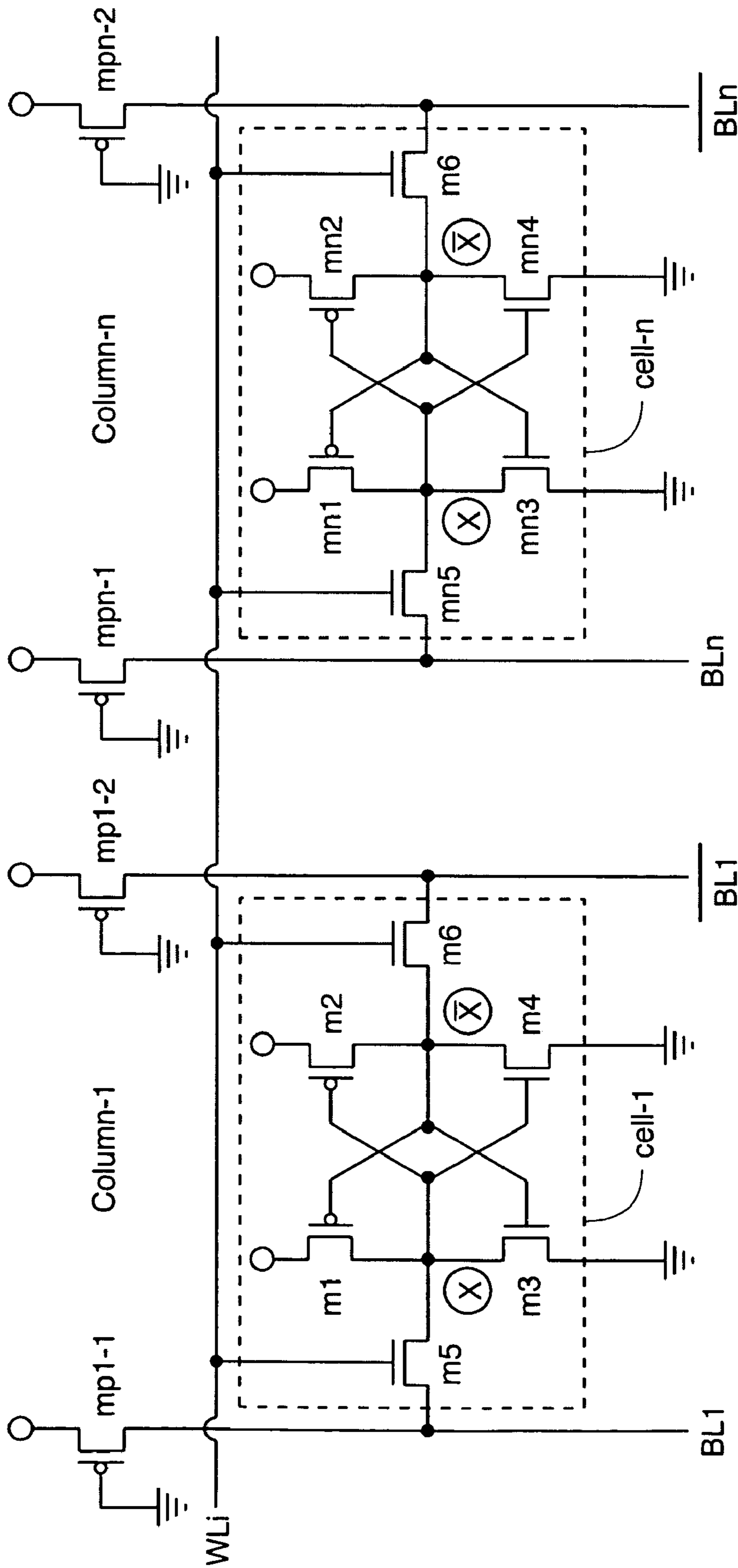
**U.S. PATENT DOCUMENTS**

5,324,992 A \* 6/1994 Maly ..... 327/172

**9 Claims, 14 Drawing Sheets**







**FIG. 1B (PRIOR ART)**



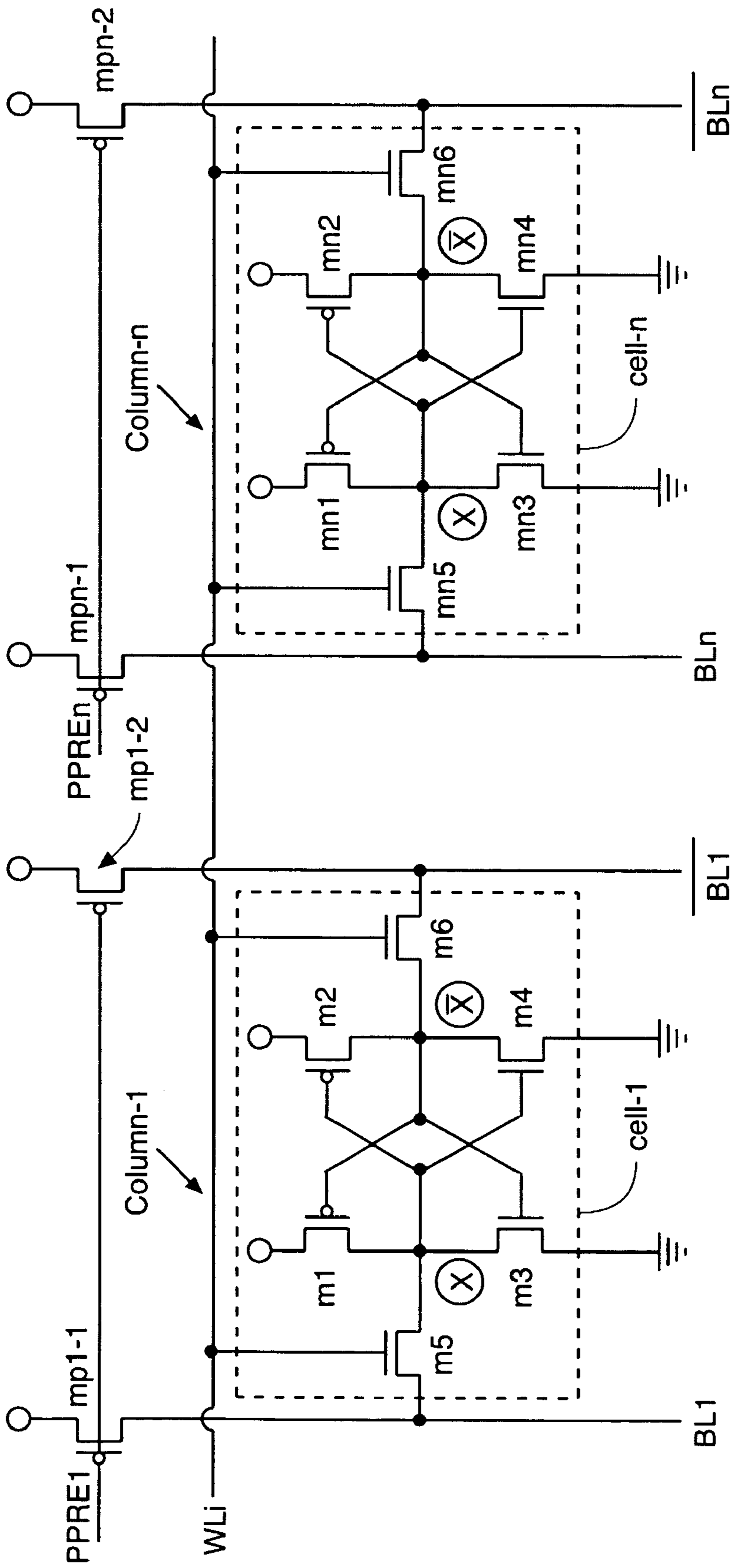
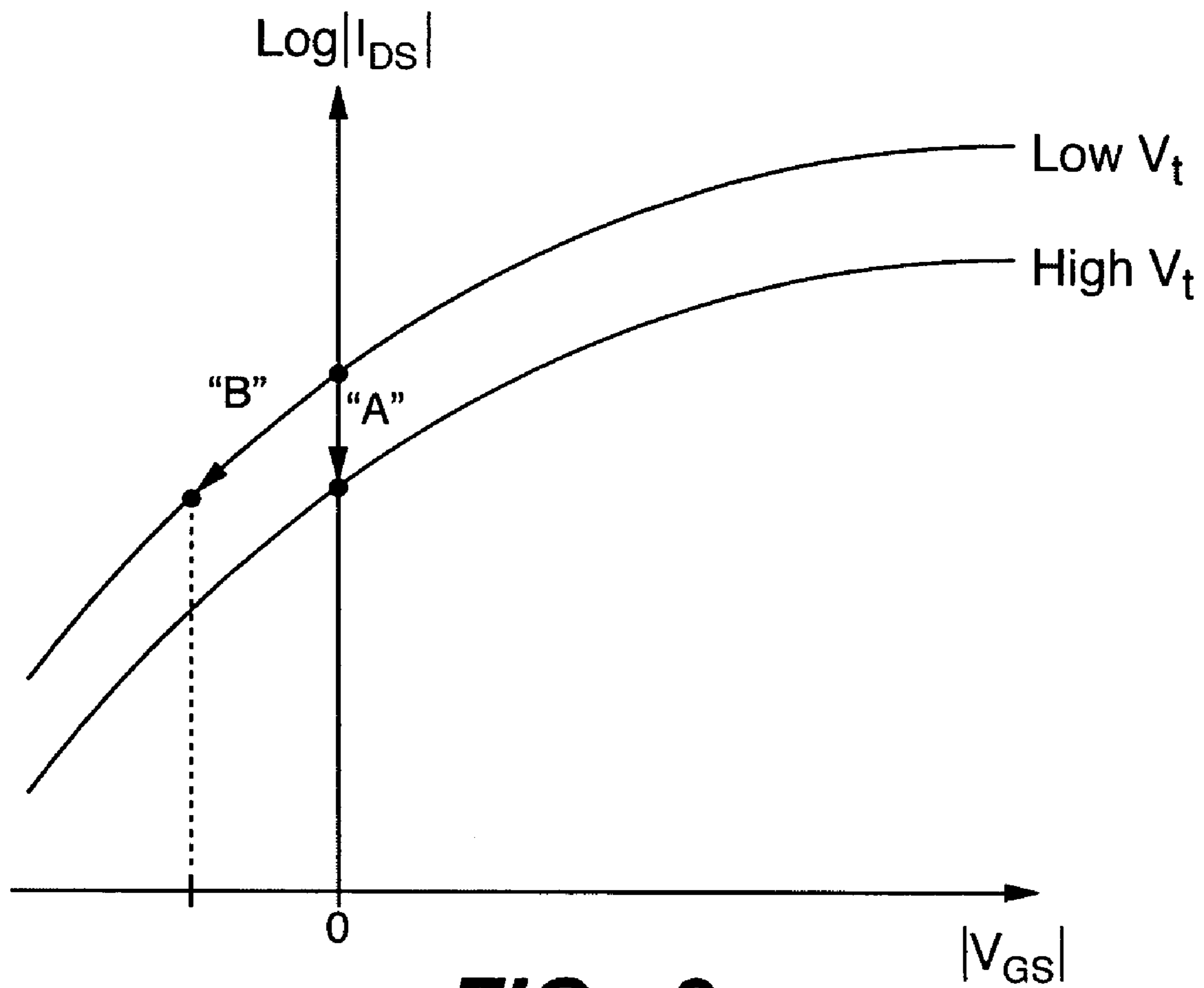
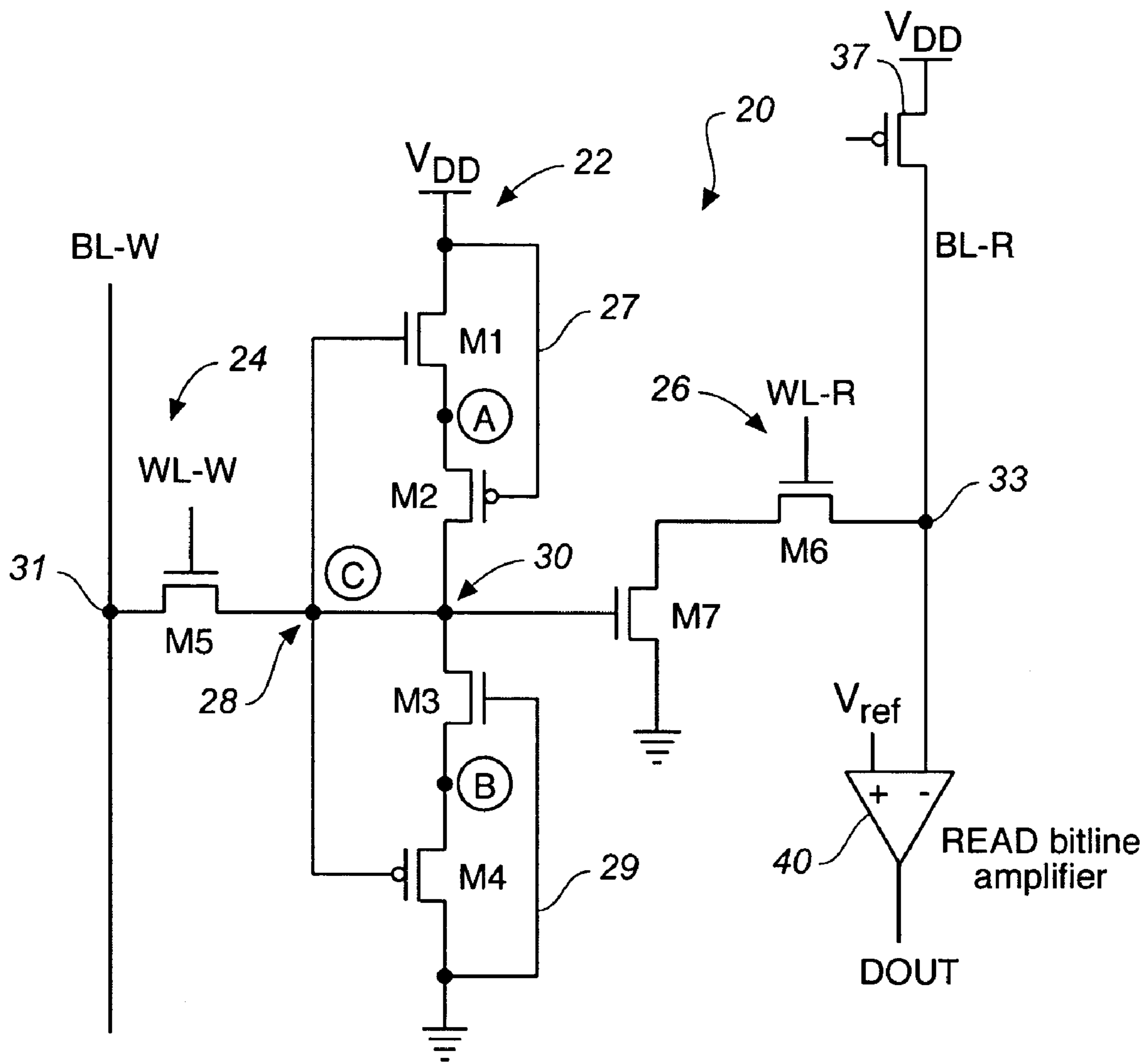


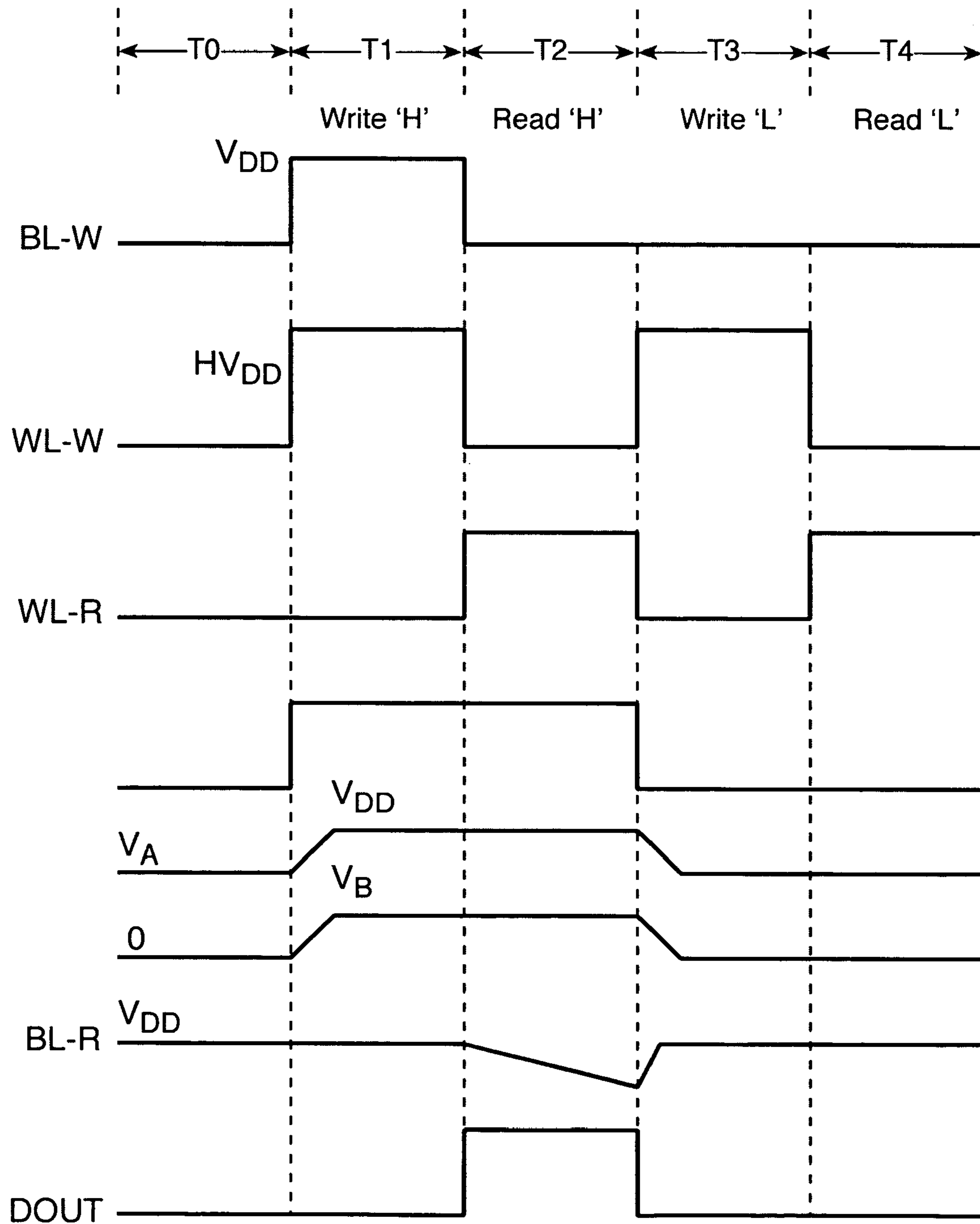
FIG. 1D (PRIOR ART)



**FIG. 2**  
**(PRIOR ART)**

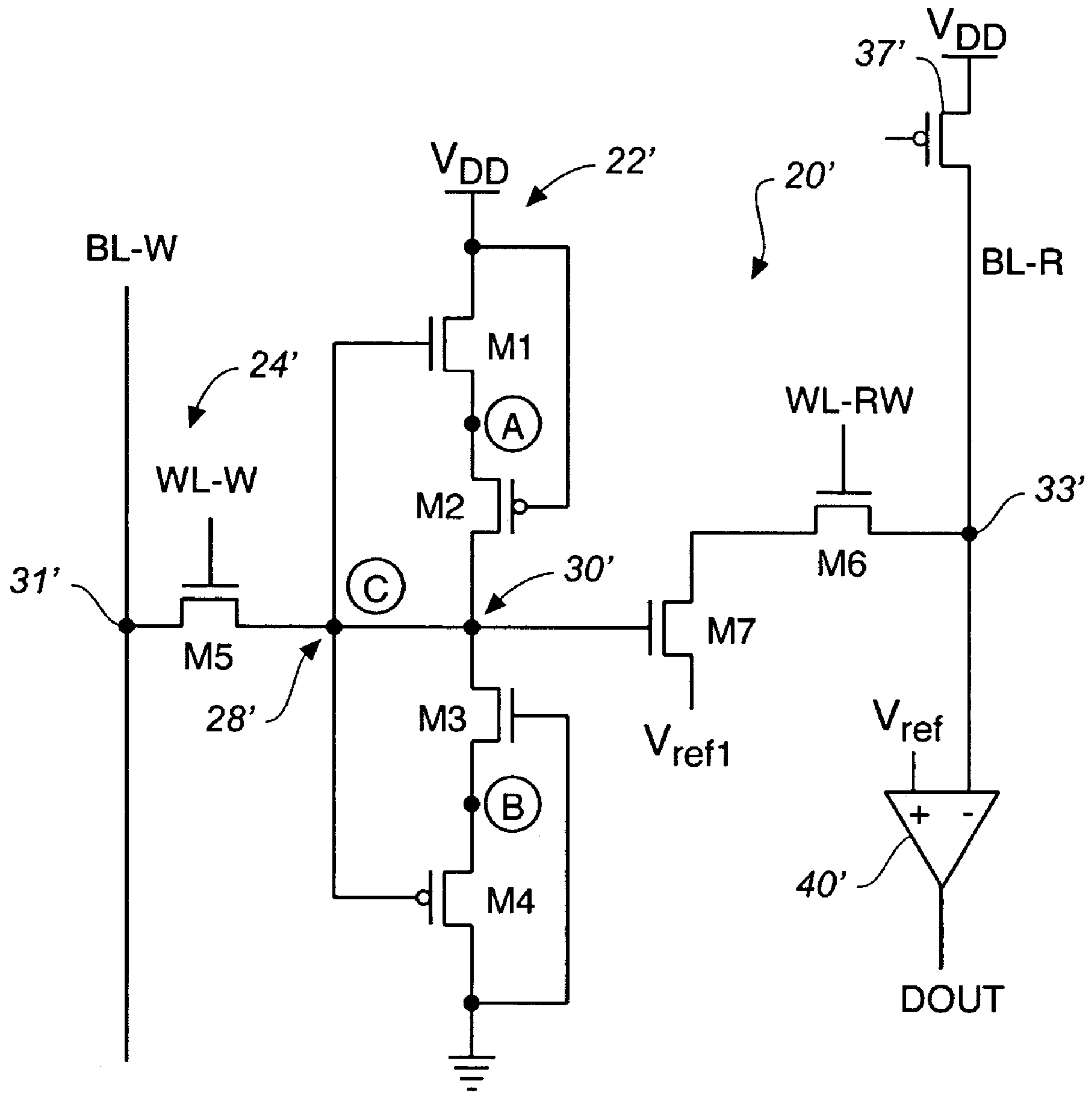


**FIG. 3**

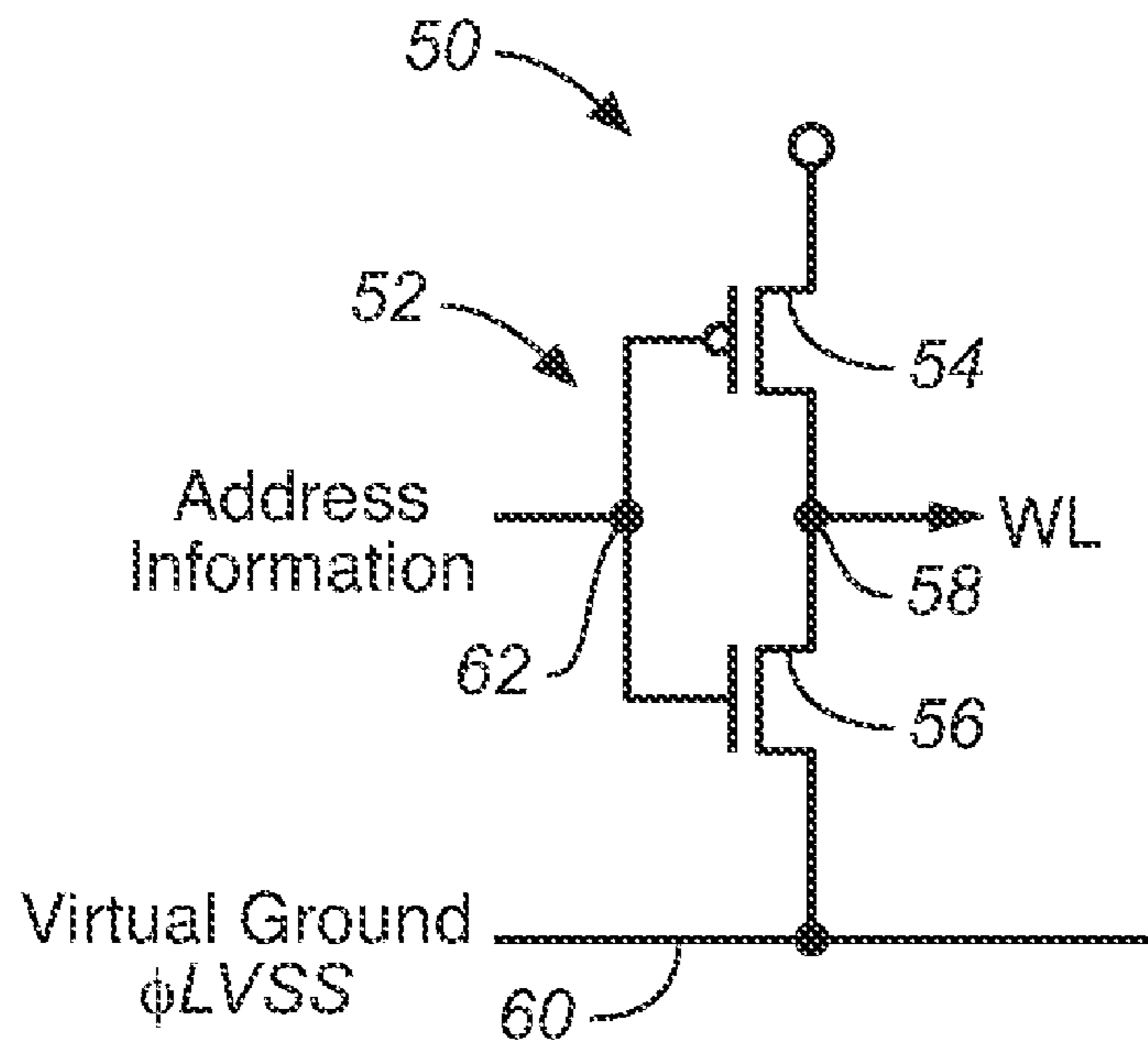


**FIG. 4**

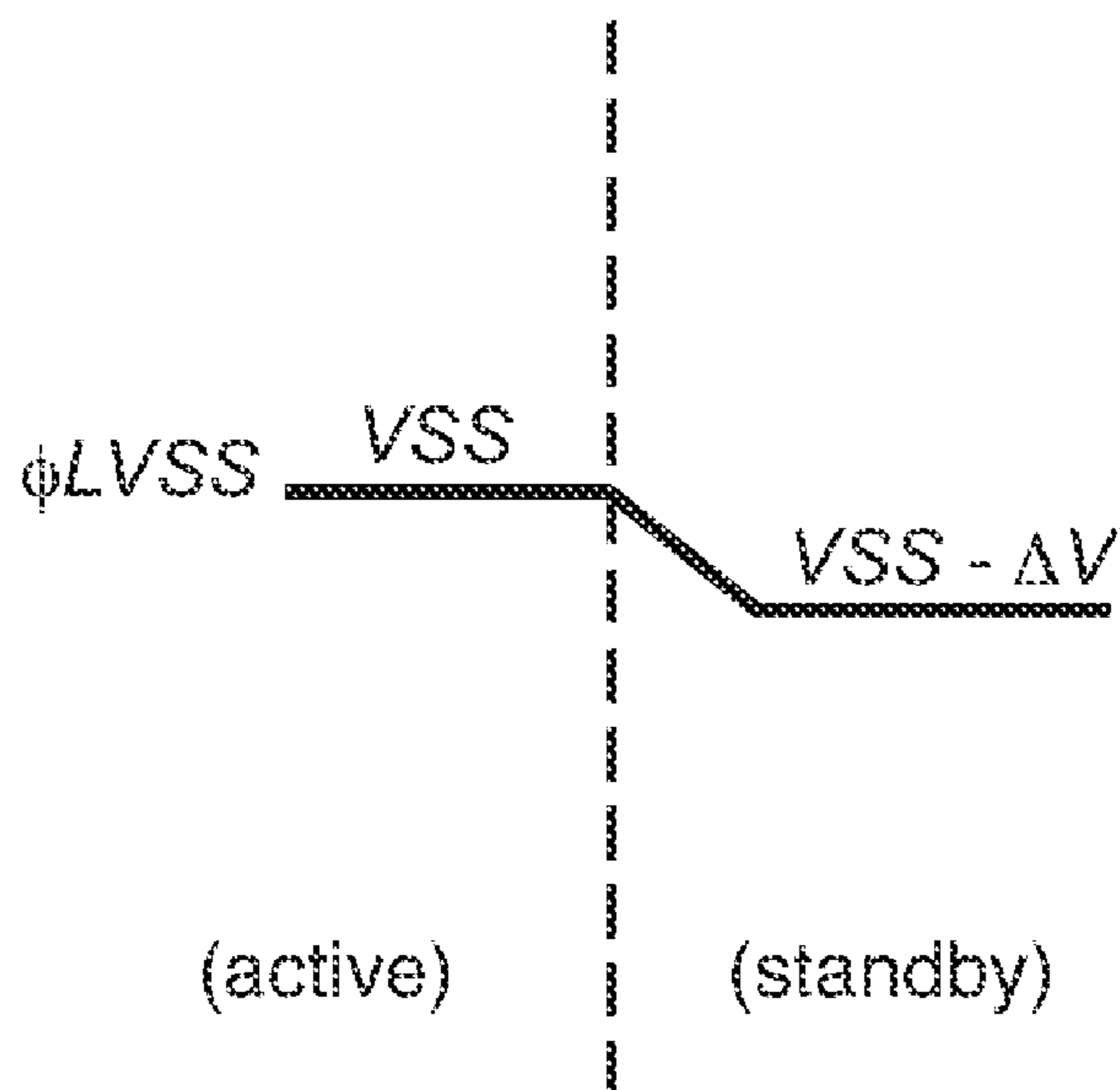




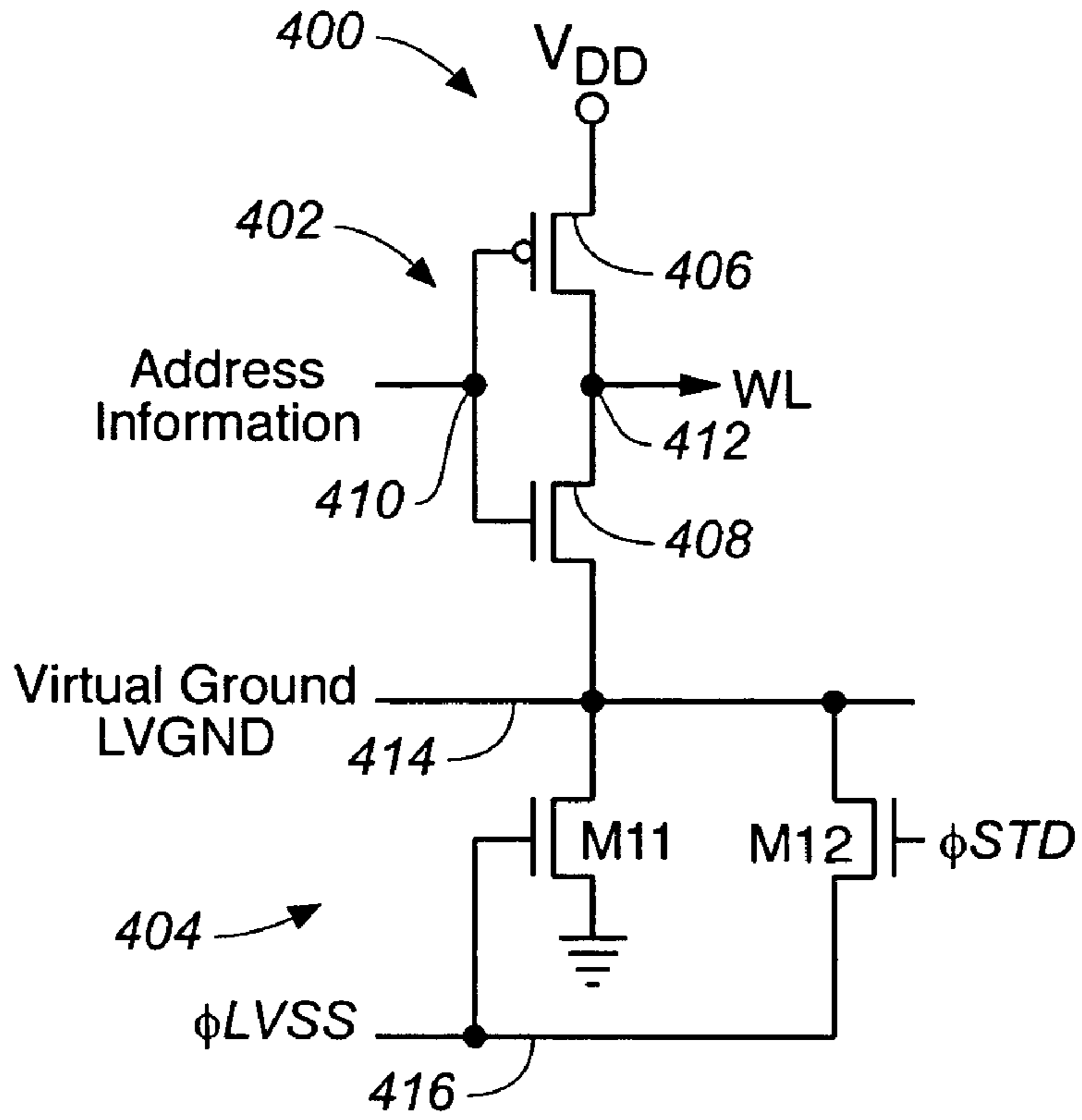
**FIG. 5**



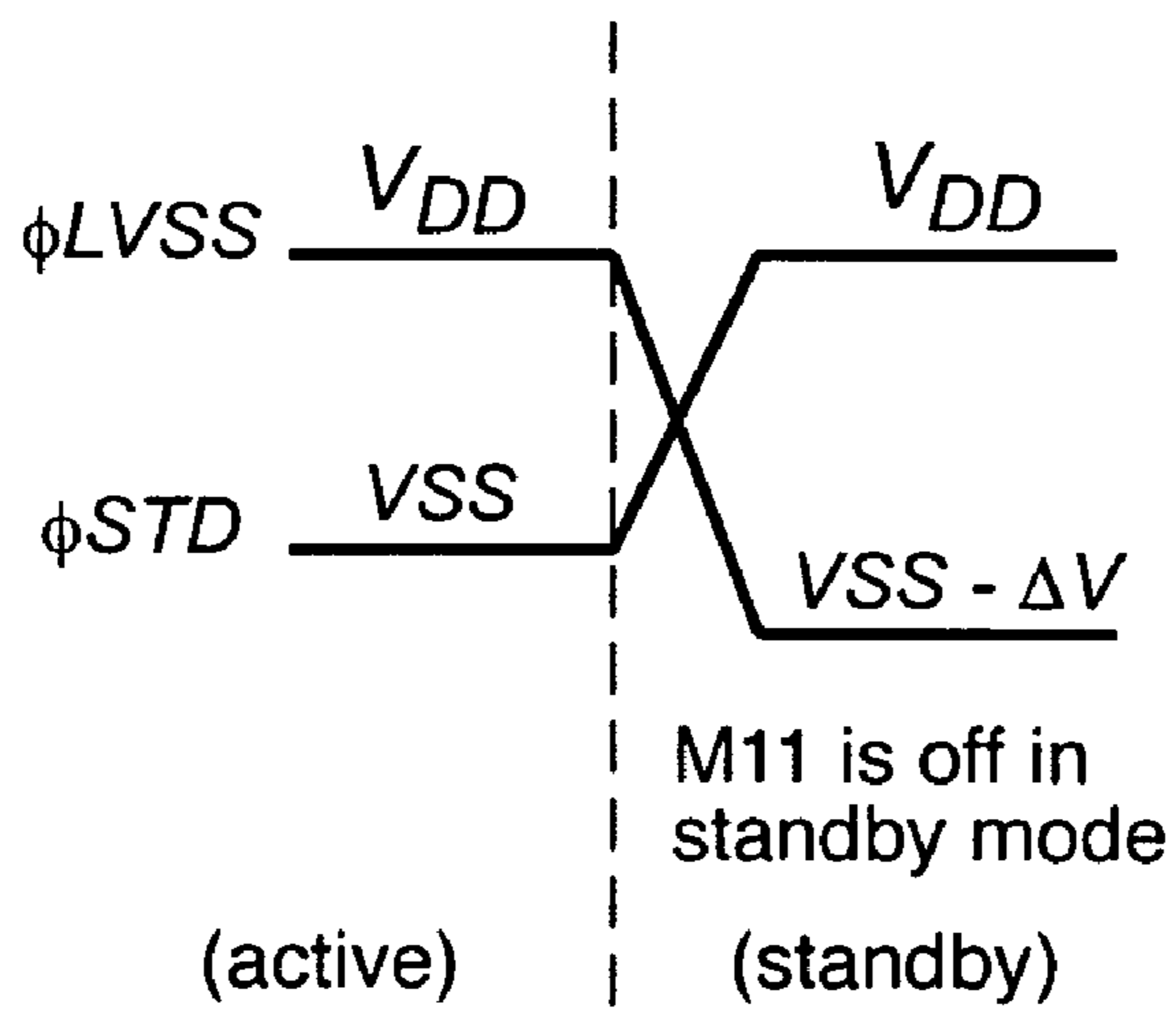
**FIG. 6A** (PRIOR ART)



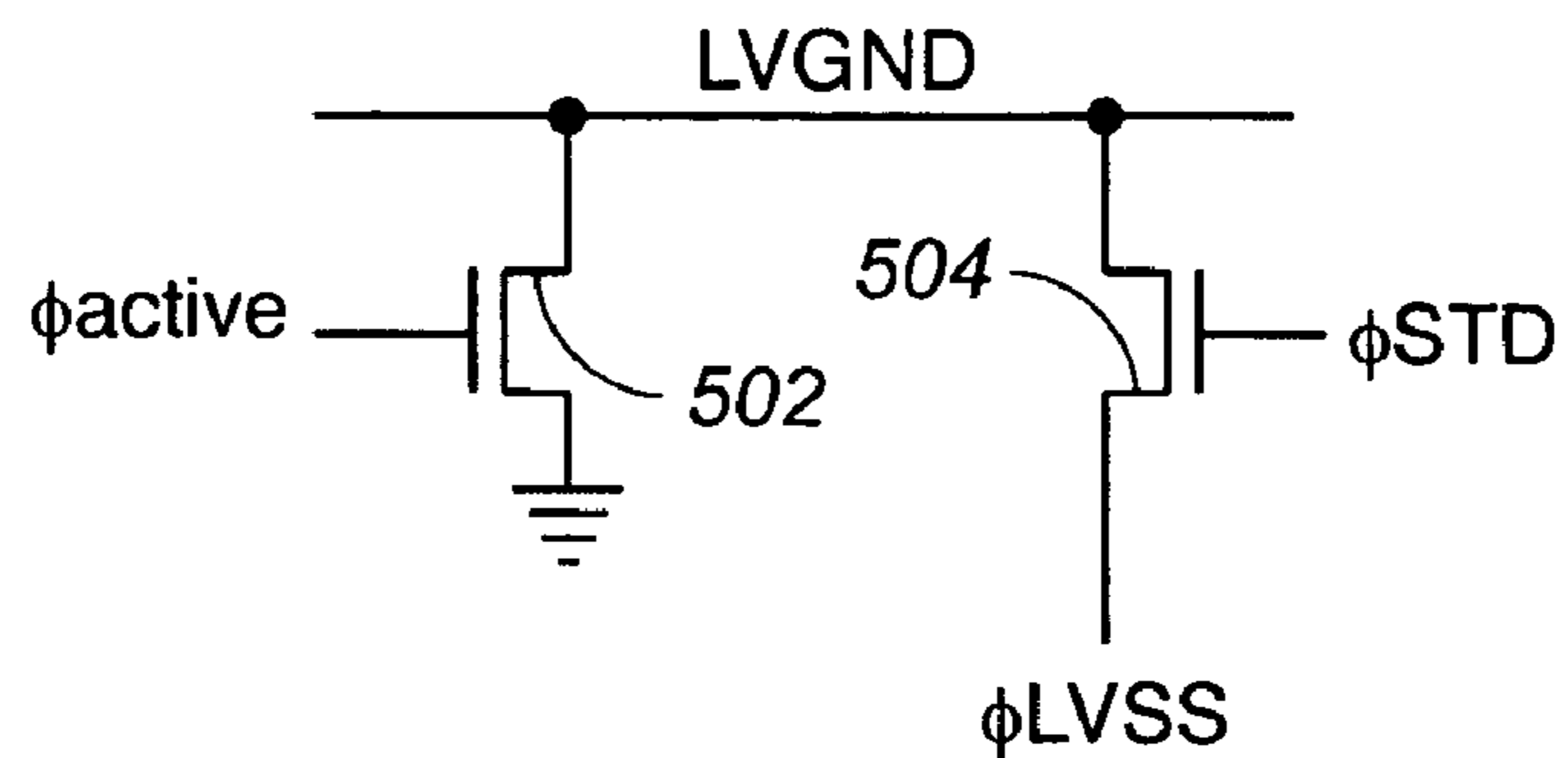
**FIG. 6B**  
(PRIOR ART)



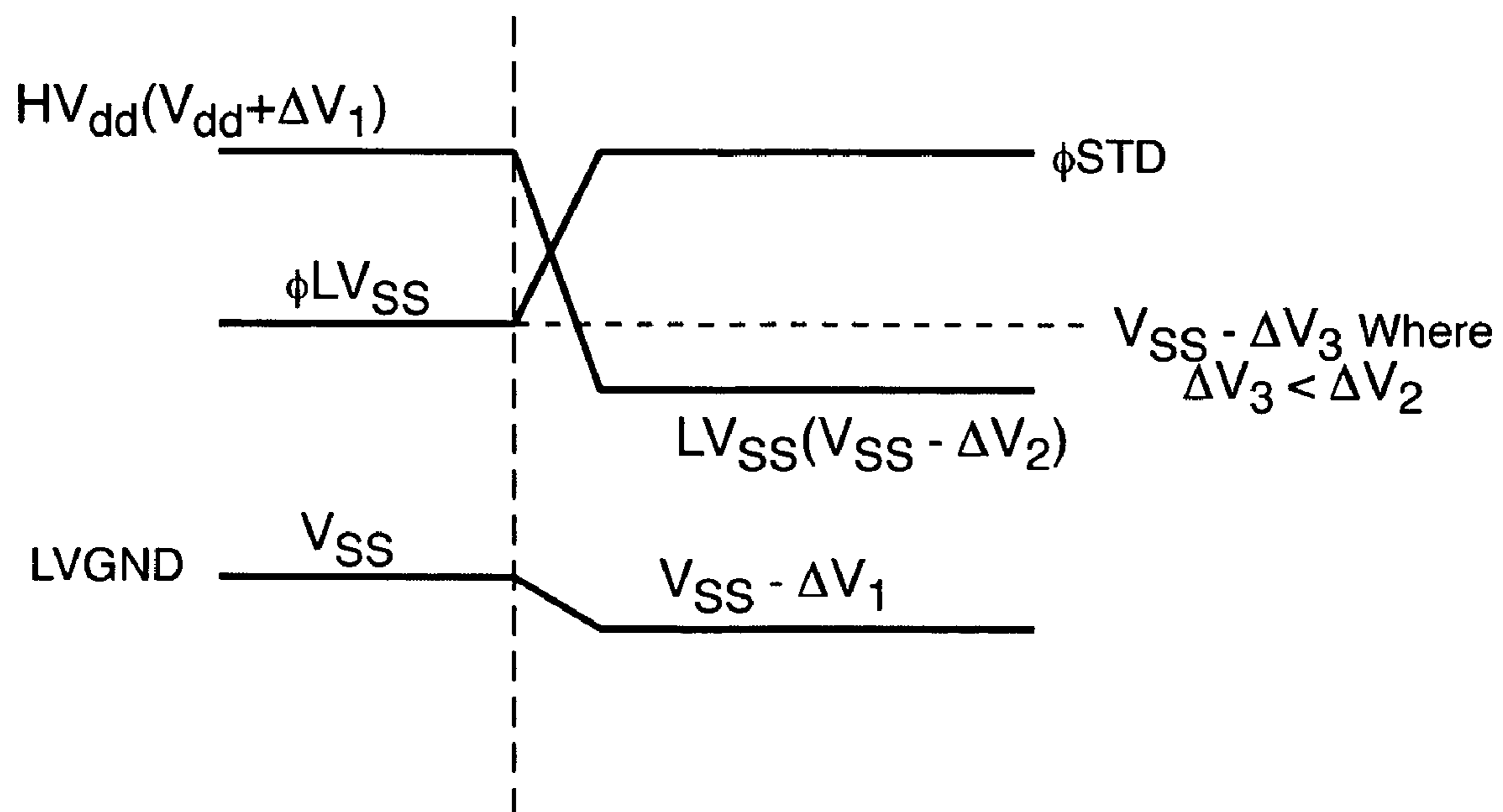
**FIG. 7A**



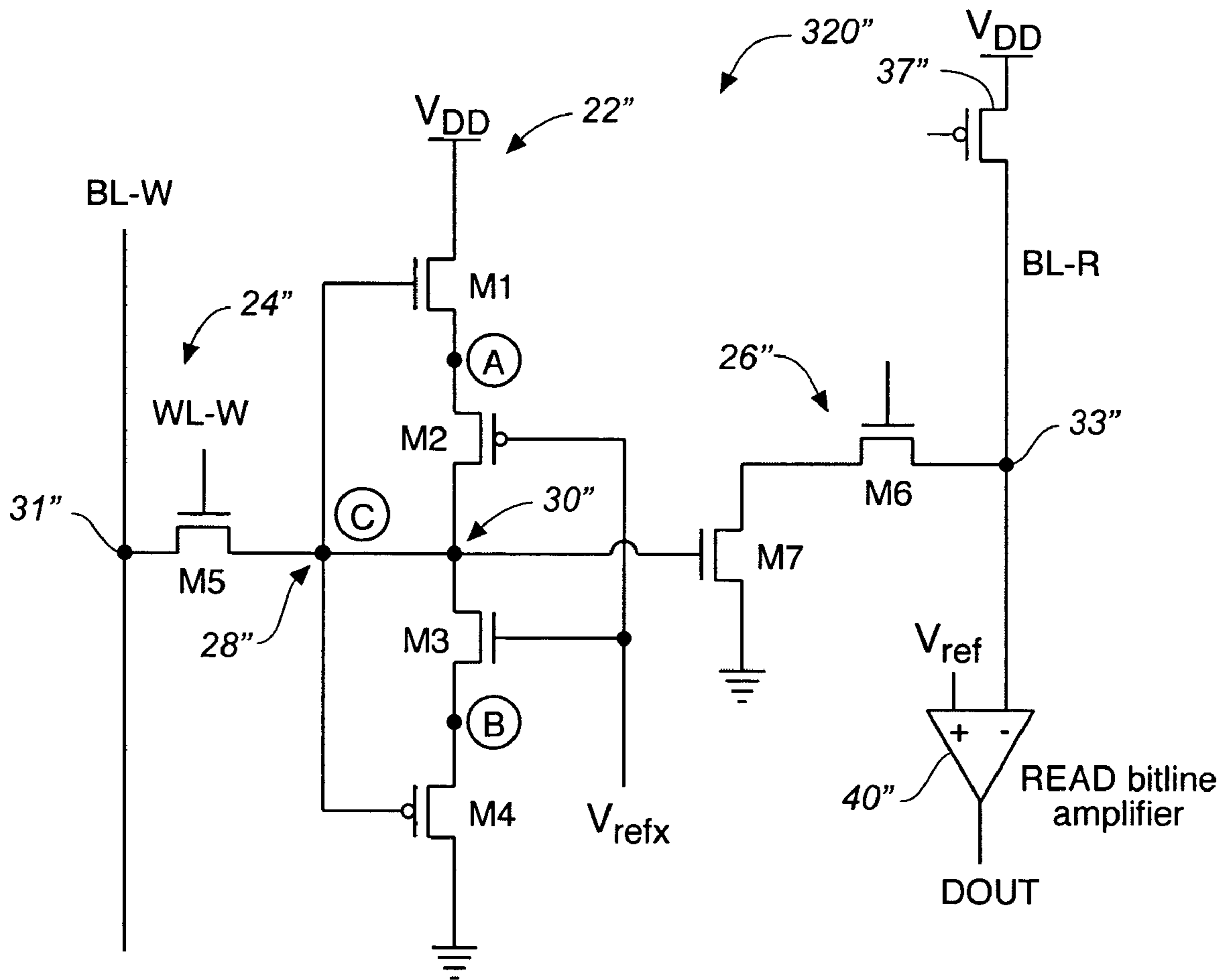
**FIG. 7B**



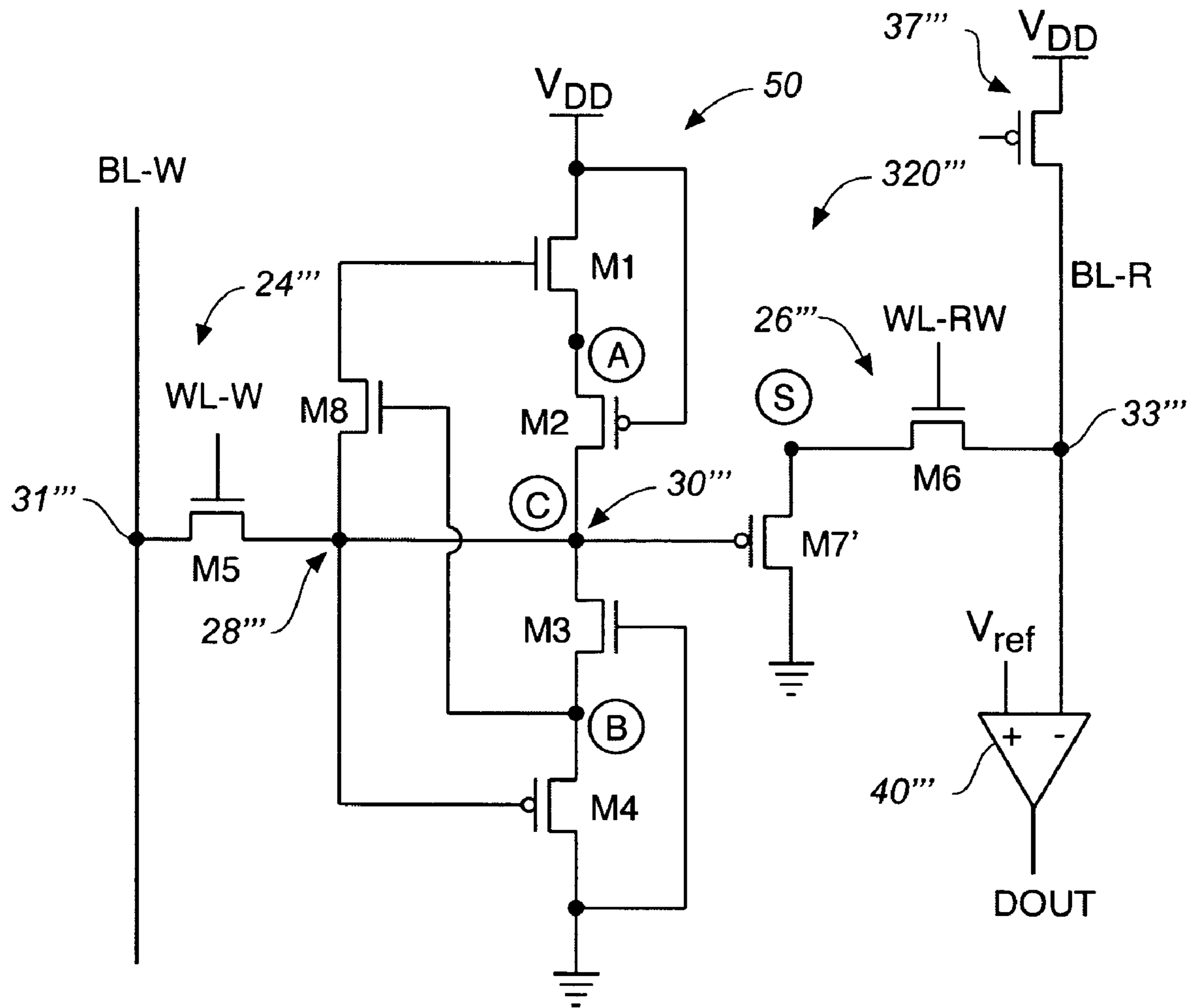
**FIG.\_8A**



**FIG.\_8B**

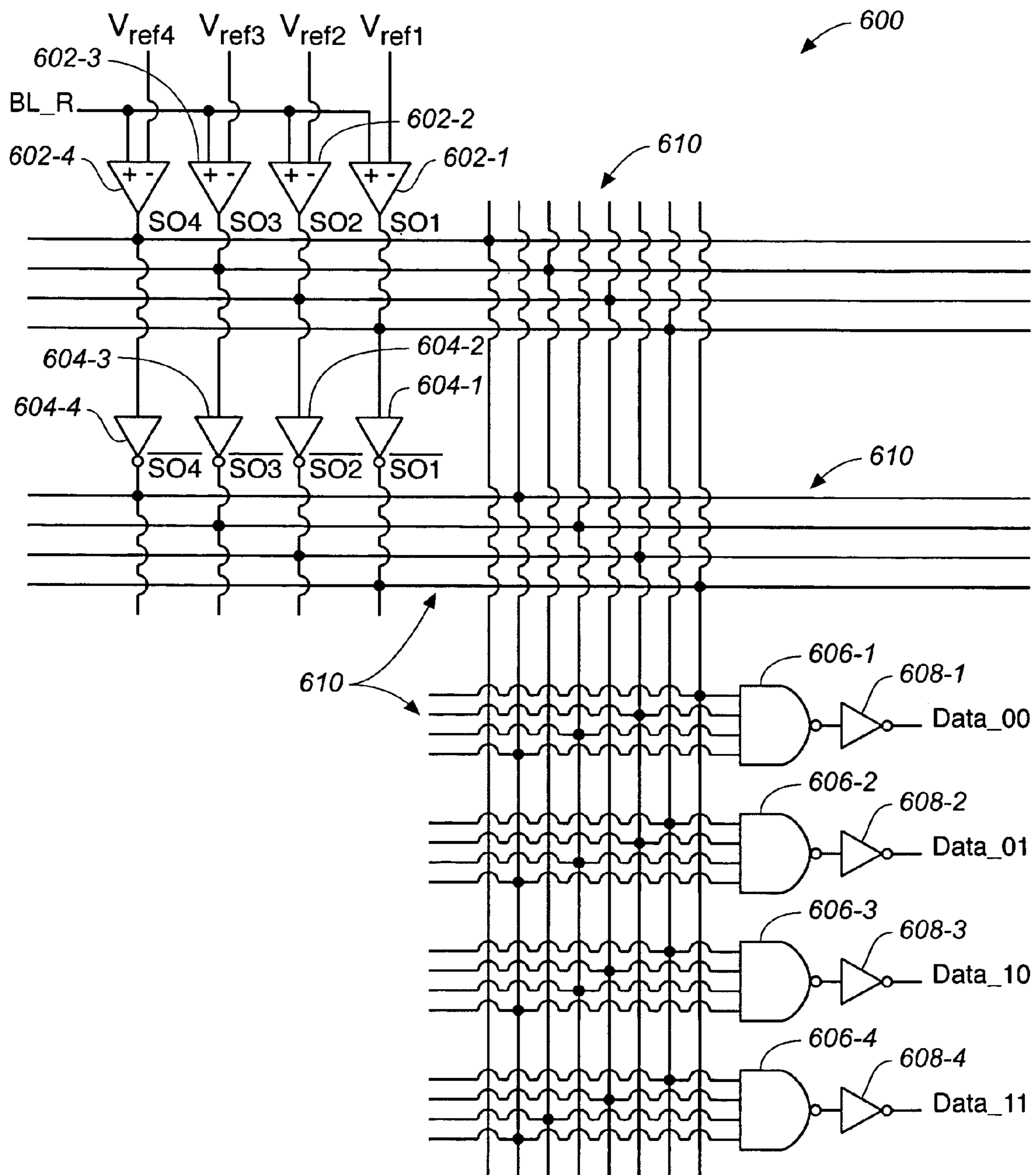


**FIG. 9**



**FIG. 10**

4-level sensing scheme



**FIG. 11**

## SELF REVERSE BIAS LOW-POWER HIGH-PERFORMANCE STORAGE CIRCUITRY AND RELATED METHODS

### CROSS-REFERENCE TO RELATED APPLICATIONS

This application is a divisional application of U.S. application Ser. No. 10/402,472, filed Mar. 27, 2003, now U.S. Pat. No. 6,992,915, and which claimed priority to and the benefit of the filing date of provisional patent application Ser. No. 60/368,392 filed Mar. 27, 2002.

### BACKGROUND OF THE INVENTION

#### 1. Field of the Invention

The invention relates to integrated circuits, and more particularly, to integrated circuits used in information storage and retrieval.

#### 2. Description of the Related Art

For over three decades the semiconductor industry has been able to take advantage of the technology scaling according to Moore's Law. A continual increase in memory chip density, and thus the on-chip memory capacity has enabled the development of new products such as portable electronic devices used for mobile computing and communications. Without high-density memory it would not have been possible to have devices such as cell phones, personal digital assistants (PDAs), palm-top computers, or even laptop computers. Power Consumption has become a significant factor in use of such portable devices. Power Consumption affects battery life, and lower power consumption leads to longer battery life. Static Random Access Memory (SRAM), has been an important component of portable devices since it consumes less power and is generally faster than dynamic RAM (DRAM) that requires periodic refresh operation to prevent loss of memory contents. Currently 16 Mb SRAMs and 256 Mb DRAMs are available on the market.

For portable device applications that require SRAM, low standby current is highly desirable so as to extend the battery lifetime. Otherwise, rapid depletion of the battery power can limit the use of portable devices and also can inconvenience users of those devices by requiring them to carry spare batteries, for example. In a present generation of portable devices, the typical standby current of a portable device is 5-10 microamperes ( $10^{-6}$  A). Ideally, the standby current would be zero, and the less the better. Standby current has several components, and one of the most significant components is due to leakage in memory cells. As the desired memory capacity for portable devices has grown, it is increasingly important to suppress the leakage current. Unfortunately, the leakage current in prior memory circuits tended to increase for each generation of technology scaling according to a physical law. It is well known that the subthreshold current of a Metal-Oxide Semiconductor (MOS) transistor increases exponentially as the device threshold voltage is scaled down as required for chip performance with a down-scaled power supply voltage. This leakage current phenomenon in an MOS transistor is described generally in the following equation:

$$I_{leakage} = K * \exp((V_{gs} - V_t) / (S / \ln 10)) (1 - \exp(-V_{ds} / V_T)) \quad (1)$$

where K is a constant that depends on the technology,  $V_{gs}$  is the gate-to-source voltage ( $=V_g - V_s$ ),  $V_t$  is the device threshold voltage, S is the subthreshold voltage swing,  $V_T$  is the thermal voltage ( $=kT/q$ ) with k denoting the Boltzman's constant. S, the subthreshold swing voltage, can be described by

$$S = (kT \ln 10) / q * (1 + C_d / C_{ox}) \quad (2)$$

Equation (1) suggests that an increase in  $V_t$  can be used to reduce the leakage current, and this approach is practiced reluctantly in VLSI design despite a speed penalty. In other words, increased  $V_t$  results in both reduced leakage current and increased signal propagation delay within the circuit. Thus, typically there has been a trade-off between a desire to minimize leakage current and a desire to maximize speed. This trade-off generally has been acceptable as long as reduced leakage current transistors with increased  $V_t$  are not in speed-critical paths. Ordinarily, transistors in speed-critical paths should have lower threshold voltages in order to ensure reduced signal propagation delay leading to increased circuit speed. However, lower threshold voltage can result in relatively heavy leakage currents in standby mode.

FIG. 1A is an illustrative circuit diagram of a known integrated circuit data storage cell of the type referred to as an SRAM cell. This prior SRAM cell includes six transistors, two transistors for access (m5, m6) and four transistors (m1, m2, m3, m4) for latching data with two cross-coupled inverters (m1-m3 pair and m2-m4 pair). Transistors m1-m4 serve as storage circuitry. In this example, the storage circuitry operates by latching data. Transistors m5 and m6 serve as access transistors for writing data to and reading data from the storage circuitry. Suppose, for example, that the stored data is logic "1." The data-storing node X is set to high ("1") and the other data-storing node X-bar is set to low ("0"). Therefore, transistors m1 and m4 are turned on while transistors m2 and m3 are turned off. Access transistors, m5 and m6, are turned on by driving wordline (WL) high and are turned off by driving wordline (WL) low. When m5 and m6 are turned on, BL is linked to node X and BL-bar is linked to X-bar.

More specifically, the integrated circuit data storage cell includes a latch circuit including first and second inverters. A first inverter includes a first high threshold voltage PMOS transistor m1 and a first high threshold voltage NMOS transistor m3 and a first data node X comprising interconnected source/drains (S/D) of the first PMOS m1 and NMOS m3 transistors. A second inverter includes a second high threshold voltage PMOS transistor m2 and a second high threshold voltage NMOS transistor m4 with a second data node X-bar comprising interconnected source/drains (S/D) of the second PMOS transistor m2 and NMOS m4 transistor. The gates of the first PMOS transistor m1 and first NMOS transistor m3 are coupled to the second data node X-bar. The gates of the second PMOS transistor m2 and the second NMOS transistor m4 are coupled to the first data node X. A first low threshold voltage access transistor m5 includes a first S/D coupled to the first data node X and to the gate of the second PMOS transistor m2 and to the gate of the second NMOS transistor m4 and includes a second S/D coupled to a first data access node A1 and includes a gate coupled to a first access control node C1. A second low threshold voltage access transistor m6 includes a first S/D coupled to the second data node X-bar and to the gate of the first PMOS transistor m1 and to the gate of the first NMOS transistor m3 and includes a second S/D coupled to a second data access node A2 and includes a gate coupled to a second access control node C2.

During a write operation, for example, when WL is high, data "1" on BL can be fed to node X by turning on access transistor m5, and at the same time, data "0" on BL-bar can be fed to node X-bar by turning on access transistor m6. The latching by m1-m3 and m2-m4 pairs enable stable storage of data "1" at node X even after the access transistors m5 and m6 are turned off with low voltage on WL line. Conversely, data "0" can be written to node X by providing logic "0" on BL while providing logic "1" on BL-bar when access transistors m5 and m6 are turned on by a high WL signal.

Conversely, during a read operation, both BL and BL-bar are pre-charged to a high voltage level, e.g.,  $V_{DD}$ . If the voltage level at node X-bar is low, then the voltage on BL-bar



will discharge through m4. If the voltage level at node X-bar is high, then the voltage on BL-bar will not discharge through m4. Instead, the voltage of BL will discharge through m3. A sense amplifier (not shown) can sense a small voltage drop on either BL or BL-bar so as to determine the voltage level stored at nodes X and X-bar and generate an output signal, e.g. high when the stored data is high or low when the stored data is low.

Unfortunately, there have been reliability problems with this earlier SRAM cell. For instance, if m3 is leaky and conducts current from node X to ground, then the charge stored at that node can be reduced, pulling down the node voltage at X, which in turn can make m2 leak some charge into node X-bar. Voltage build up at node X-bar can in turn promote more leakage current through m3 potentially causing a transition to a new erroneous locked state. Thus, current leakage can result in reliability problems by causing erroneous data storage.

Another problem with the prior SRAM structure of FIG. 1A is that the voltage of a cell node can be influenced by bit line voltages during a read operation. For example, assume that the voltages of X and X-bar are high and low respectively, and BL and BL-bar are precharged at  $V_{DD}$ . When WL is enabled, m5 and m6 become to turned on. Node X at the junction of m1 and m3 is coupled to BL, and node X-bar at the junction of m2 and m4 is coupled to BL-bar. Since BL-bar is precharged to VDD and the level of X-bar is low (e.g., VSS), the voltage level on BL-bar can influence the voltage level at node X-bar. One approach to reducing the influence that the voltage on BL-bar has on node X-bar is to increase the impedance of m6. However, increasing the impedance of m6 also slows down the read speed. Thus, there is a tradeoff between circuit stability and read speed.

In addition, leakage currents in standby mode can cause draining of the battery. FIG. 2 is an illustrative drawing of an I-V characteristics of a MOS transistor for two different threshold voltages (low  $V_t$  and high  $V_t$ ). As explained by equation (1), the higher the threshold voltage, the lower the leakage current ( $I_{ds}$ ) in magnitude. Thus, design option "A" would use high  $V_t$  to lower the leakage current, but this would cause speed degradation since the signal propagation delay increases as  $V_t$  increases for a fixed  $V_{gs}$  ( $<V_{DD}$ ) swing. It is known that the propagation delay driven by a MOS transistor is inversely proportional to  $(V_{gs}-V_t)$ . Thus, for a given  $V_{gs}$ , there is more delay at higher  $V_t$ . By comparison, design option "B" would use low  $V_t$  to increase speed by reducing signal propagation delay but leakage current can be reduced by making the transistor reverse-biased when the transistor is turned-off.

Although the fact that leakage current can be reduced when a transistor is reverse biased has been known, there still exists a need for an integrated circuit data storage cell that suppresses leakage current without performance degradation and reliability issues. In particular, there has been a need for an SRAM cell that reduces power consumption without sacrificing high speed performance

There also has been a need for increased memory capacity within a given chip area. In order to store more information within a given area of a chip, the individual storage cell area should be small. For this reason, in the past, layout experts often did data storage cell layout manually. Even a tiny saving in the unit cell area can lead to significantly increased storage capacity within an overall chip area, especially when many data storage cells are used repeatedly on a chip. One approach that has been considered for increasing information storage capacity is to store more than one bit of data in a single memory cell. If two data bits can be stored in one unit cell,

then effectively the memory capacity can be doubled for the same chip area. Moreover, the chip area for a given data storage capacity could be reduced, thus increasing the production yield. There has been a need for a memory cell architecture that allows a simple approach to implementing multiple-bit storage in a single memory cell.

In addition, there has existed a need to reduce power consumption due to precharging of bit lines for read operations and to reduce chip area occupied by precharge circuitry. FIGS. 1B-1D are illustrative circuit diagrams showing the known cell of FIG. 1A coupled in a typical SRAM array structure in which each array has m-rows (WL) and n-columns (BL and BL-bar) of cells. Each column includes a bit line pair BL and BL-bar. Each away of FIGS. 1B-1D has a different known precharge circuitry configuration. More particularly, FIGS. 1B-1D show a first cell, cell-1 in a first column column-1 and a last cell, cell-n of a last column column-n of row m. Specifically, cell-1 and cell-n include respective input transistors m5 and m6 with gates coupled to WLm. Respective S/D terminals of m5 and m6 input transistors of cell-1 are respectively coupled to BL1 and BL1-bar. Respective S/D terminals of m5 and m6 input transistors are respectively coupled to BLn and BLn-bar.

FIG. 1B shows a first precharge circuitry configuration in which, at the end of each bit line, a precharge transistor, e.g., PMOS transistors mp1-1, mp1-2, mpn-1 and mpn-2, is placed to set the bit line (BL) and bit line bar (BL-bar) voltages at a certain level. For example, the notation "mp1-1" indicates, row-m, precharge, column-1, coupled to first bit line (BL1) of cell. For example, the notation "mpn-2" indicates, row-m, precharge, column-n, coupled to second bit line (BLn-bar) of the cell.

In the precharge circuitry configuration of FIG. 1B, all BL lines and all BL-bar lines are precharged to  $V_{DD}$  through mp1-1, mp1-2, mpn-1 and mpn-2. The gates of these transistors are coupled to a power supply level, e.g.,  $V_{SS}$  in this example, and thus, the precharge transistors are always turned on, and the BL lines and BL-bar lines are continually precharged to  $V_{DD}$ .

FIG. 1C shows a second precharge circuitry configuration similar to that of FIG. 1B. However in the configuration of FIG. 1C the precharge transistors are controlled by a precharge control signal PPRE.

FIG. 1D shows a third precharge circuitry configuration in which each bit line pair has a dedicated control signal, e.g., PPRE1 for BL1 and BL1-bar and PPREn for BLn and BLn-bar. These dedicated control signals permit selective precharging of bit pairs. Selective precharging can reduce power consumption since current paths not involved in a read operation are not unnecessarily precharged. Moreover, cells in a given SRAM array can be divided into groups, and different respective precharge signals can be used to control precharging of different respective groups of cells. For example, assuming that there are 128 cells in an SRAM array, and that only 16 bits of cell data are read at a time, then the cells of the array can be divided into 8 groups, and each group can have a different precharge control signal.

While a precharge circuitry configuration such as that of FIG. 1D can reduce precharge-related power consumption, there has existed a need for further improvements in precharge-related power consumption and for precharge circuitry that occupies less chip area.

The present invention meets these needs.

## 5

## SUMMARY OF THE INVENTION

One aspect of the invention, for example, provides an integrated circuit storage device. The device includes first NMOS and PMOS transistors and second NMOS and PMOS transistors. A first source/drain of the first NMOS transistor serves as a first bias node. A first source/drain of the second PMOS transistor serves as a second bias node. A gate of the first PMOS transistor serves as a third bias node. A gate of the second NMOS transistor serves as a fourth bias node. A junction of a second source/drain of the first NMOS transistor and a first source/drain of the first PMOS transistor serves as a pull-up node. A junction of a second source/drain of the second PMOS transistor and a first source/drain of the second NMOS transistor serves as a pull-down node. The storage device also includes an input node, a storage node and an output node. The storage node is at a junction of a second source/drain of the first PMOS transistor and a second source/drain of the second NMOS transistor. An input switch controls transmission of an input data value from the input node to a gate of the first NMOS transistor and to a gate of the second PMOS transistor. An output switch controls transmission of a stored data value from the storage node to the output node.

Another aspect of the invention, for example, provides an integrated circuit multi-voltage level storage device. The device includes first NMOS and PMOS transistors and second NMOS and PMOS transistors. A first source/drain of the first NMOS transistor serves as a first bias node. A first source/drain of the second PMOS transistor serves as a second bias node. A gate of the first PMOS transistor serves as a third bias node. A gate of the second NMOS transistor serves as a fourth bias node. A junction of a second source/drain of the first NMOS transistor and a first source/drain of the first PMOS transistor serves as a pull-up node. A junction of a second source/drain of the second PMOS transistor and a first source/drain of the second NMOS transistor serves as a pull-down node. The storage device also includes an input node, a storage node and an output node. The storage node is at a junction of a second source/drain of the first PMOS transistor and a second source/drain of the second NMOS transistor. An input switch controls transmission of a data input signal that can have any of multiple prescribed input signal voltage levels from the input node to a gate of the first NMOS transistor and to a gate of the second PMOS transistor. Limiting circuitry limits the storage node voltage to a prescribed storage node voltage level determined by a most recent data input signal voltage level. An output switch controls transmission from the storage node to the data output node, of a data output signal indicating the determined storage node voltage level.

Another aspect of the invention provides improved dynamic bias circuitry and methods.

These and other features and advantages of the invention will be appreciated from the following detailed description of embodiments of the invention and through reference to the illustrative drawings.

## BRIEF DESCRIPTION OF THE DRAWINGS

FIG. 1A is an illustrative circuit diagram of a known integrated circuit data storage cell of the type generally referred to as an SRAM cell. FIGS. 1B-1D are three illustrative circuit diagrams of the known cell of FIG. 1A in a typical SRAM array structure with three different precharge circuitry configurations.

## 6

FIG. 2 is an illustrative drawing of an I-V characteristics of a MOS transistor for two different threshold voltages (low  $V_t$  and high  $V_t$ ).

FIG. 3 is an illustrative circuit diagram of an integrated circuit data storage cell in accordance with a first embodiment of the invention.

FIG. 4 is an illustrative timing diagram used to explain the operation of the circuits of FIGS. 3, 5, 9 and 10.

FIG. 5 is an illustrative diagram of an integrated circuit data storage cell in accordance with a second embodiment of the invention.

FIG. 6A is an illustrative drawing of a conventional word line driver circuit that can be used with data storage cell circuitry of embodiments of the present invention.

FIG. 6B is an illustrative drawing of a virtual ground signal applied to the driver of FIG. 6A during active and standby modes of operation.

FIG. 7A is a circuit diagram of a word line driver circuit in accordance with one aspect of the invention.

FIG. 7B is a signal diagram used to explain the operation of the word line driver circuit of FIG. 7A in active and standby modes.

FIG. 8A shows an alternative embodiment of only a bias circuitry portion word line driver circuit in accordance with one aspect of the invention.

FIG. 8B is a signal diagram used to explain the operation of the bias circuitry of FIG. 8A.

FIG. 9 is an illustrative diagram of an integrated circuit data storage cell in accordance with a third embodiment of the invention.

FIG. 10 is an illustrative diagram of multi-state storage circuitry in accordance with a fourth embodiment of the invention.

FIG. 11 is an illustrative drawing of a multi-level sense amplifier that can be used with the embodiment of FIG. 12.

## DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENT

The present invention provides novel integrated circuitry that can exhibit high-performance (high-speed) operation in an active mode, can significantly suppress subthreshold leakage current in a standby mode, and can operate with a relatively low (less than 1V) supply voltage. The following description is presented to enable any person skilled in the art to make and use the invention. The embodiments of the invention are described in the context of particular applications and their requirements. These descriptions of specific applications are provided only as examples. Various modifications to the preferred embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments and applications without departing from the spirit and scope of the invention. Thus, the present invention is not intended to be limited to the embodiments shown, but is to be accorded the widest scope consistent with the principles and features disclosed herein.

FIG. 3 is an illustrative circuit diagram of an integrated circuit data storage cell 20 in accordance with a first embodiment of the invention. The storage cell 20 includes storage circuitry 22, input switch control circuitry 24 which inputs information for storage by the storage circuitry 22 and output switch control circuitry 26 which outputs information stored by the storage cell. The storage circuitry 22 includes transistors M1, M2, M3 and M4. The input switch control circuitry 24 includes input transistor M5. The output switch control circuitry 26 includes output transistors M6 and M7.

The storage circuitry **22** includes a first NMOS transistor **M1** with a drain coupled to a supply voltage  $V_{DD}$  and also includes a first PMOS transistor **M4** with a drain coupled to an effective ground voltage  $V_{SS}$ . The storage circuitry **22** also includes a second PMOS transistor **M2** with a source coupled to the source of the first NMOS transistor **M1** and also includes a second NMOS transistor **M3** with a source coupled to a source of the first PMOS transistor **M4**. A junction of the sources of the first NMOS transistor and the first PMOS transistor comprises a pull-up node labeled "A". A junction of the sources of the second NMOS transistor and the second PMOS transistor comprises a pull-down transistor labeled node "B". A gate of the second PMOS transistor **M2** is coupled to the supply voltage  $V_{DD}$ . More specifically, conductor node **27** couples the supply voltage to a gate of PMOS transistor **M2** so that the supply voltage biases the gate of **M2**. A gate of the second NMOS transistor **M3** is coupled to the effective ground voltage  $V_{SS}$ . Moreover, conductor node **29** couples the effective ground voltage to a gate of NMOS transistor **M3** so that the effective ground voltage biases the gate of **M3**. Gates of the first NMOS transistor **M1** and the first PMOS transistor **M4** are coupled via node **28** to the storage circuitry **22**. A drain of the second PMOS transistor **M2** and a drain of the second NMOS transistor **M3** are coupled to provide a storage node **30**. Actually, in a present embodiment, nodes **28** and **30** behave as a single node **28/30** of the storage circuitry **22**. In order to store a digital signal value, transistors **M1-M4** cooperate to maintain storage node **28/30** at a logic level of the logic value provided on input node **28** during the most recent data write cycle.

The input switch control circuitry **24** includes input transistor **M5** with a first source/drain (S/D) terminal coupled to an input node **31** and with a second S/D terminal coupled to node **28/30**. In the first embodiment, input node **31** is part of an input signal source comprising a bit line-write (BL-W) conductor line. A gate of transistor **M5** is coupled to receive a write control signal. In the first embodiment, the write control signal is referred to as a word line-write (WL-W) signal. In the first embodiment, transistor **M5** is an NMOS device.

The output switch control circuitry **26** includes output transistors **M6** and **M7**. A first S/D terminal of **M7** is coupled to the effective ground  $V_{SS}$ , and a second S/D terminal of **M7** is coupled to a first S/D terminal of **M6**. The effective ground serves as a discharge path during reading of logic level 1 signals, as explained below. A second S/D terminal of **M6** is coupled to an output node **33**. In the first embodiment, the output node **33** is part of a bit line-read (BL-R) conductor line. A gate of transistor **M7** is coupled to the storage node **30** of the storage circuitry **22**. A gate of transistor **M6** is coupled to receive a read control signal. In the first embodiment, the read control signal is referred to a word line-read (WL-R) signal. In the first embodiment, transistors **M6** and **M7** are NMOS devices.

Precharge circuitry is coupled to provide a precharge voltage to BL-R. In a present embodiment, precharge circuitry for storage cell **20** includes PMOS transistor **37** with one S/D node coupled to a  $V_{DD}$  supply and with another S/D node coupled to the BL-R bit line and with a gate coupled to receive a precharge control signal. Alternatively, an NMOS precharge transistor may be employed. In one embodiment, the precharge control signal is a constant value signal that maintains transistor **37** turned on continuously. In another embodiment, the precharge control signal turns on transistor **37** only during read operations.

It will be understood that in the embodiment of FIG. 3, only one precharge line per storage cell is employed, since only one bit line (BL-R) is used to read stored data. Only one bit

line per storage cell (BL-R) has to be precharged to effect a read operation. Thus, less chip area is required and less power is consumed in connection with precharging of bit lines for read operations.

Transistors **M1-M4** are preferably depletion transistors or equivalently 'leaky' enhancement transistors. As used herein, the term depletion transistor shall include 'leaky' enhancement transistors. In a present embodiment, **M1-M4** are low  $V_t$  transistors. A MOS transistor is a depletion transistor if the transistor is turned on even when the gate-to-source voltage ( $V_{gs}$ ) is 0V. An NMOS depletion transistor can be produced by implanting n-type impurities in the transistor's channel region such that strong channel conduction can be achieved even with  $V_{gs}=0V$ . Similarly, a PMOS depletion transistor can be produced by implanting p-type impurities in the transistor's channel region such that strong channel conduction can be achieved even with  $V_{gs}=0V$ . A 'leaky' enhancement transistor as the term is used herein means an enhancement transistor having insufficient current driving capability to change the state of a node within a given timing constraint but having a larger current than the junction leakage current of the node. Typically, transistors having very low threshold voltages are depletion transistors while transistors having higher threshold voltages are enhancement transistors.

Transistors **M5, M6** and **M7** of the first embodiment **20** are 'ordinary' enhancement transistors. **M5-M7** have higher threshold voltages than **M1-M4**. To improve read speed, **M6** and **M7** are recommended to have lower threshold voltages than **M5**, although it is not required that **M6** and **M7** have a lower threshold voltage than **M5**. As used herein, the term 'ordinary' enhancement transistor means that the transistor is in an off state when a magnitude of a gate-to-source voltage of the transistor is zero.

In operation, during a write cycle, a logical 0 or a logical 1 digital information signal provided on input node **31** and on the BL-W is passed by transistor **M5**, causing the storage circuitry **22** to store a digital value representing the information signal. During the write cycle, transistor **M6** de-couples the storage cell **22** from the output node **33** and BL-R. Following a write cycle, transistors **M5** and **M6** turn off, isolating node **28/30** from the input node **31** and the output node **33**. Following the write cycle, the storage circuitry **22** stores a logic value most recently written from the BL-W line via transistor **M5**.

In operation, during a read cycle, output transistors **M6** and **M7** cooperate to effect a signal on the output node **33** and on the BL-R output that is indicative of the logic value currently stored at storage node **28/30** by the storage circuitry **22**. Output transistor **M6** is turned on during a read cycle. However, the on/off state of the output transistor **M7** during a read cycle depends upon the logic level maintained at storage node **28/30** by the storage circuitry **22**. More specifically, for example, in one embodiment of the invention when a stored logic level 0 signal is read from node **28/30** of the storage circuitry **22** there is no discharge to ground of a pre-charge voltage on the BL-R bit line through transistor **M7** because **M7** is turned off due to the provision of a logical 0 signal to its gate by the storage circuitry output terminal. Conversely, for example, when a stored logic level 1 signal is read from node **28/30** the storage circuitry **22** there is a discharge to ground of a pre-charge voltage on output node **33** of the BL-R bit line, through transistor **M7**, because **M7** is turned on due to the provision of a logical 1 signal to its gate by the storage node **28/30**.

In the course of the read cycle, a sense amplifier **40** senses whether or not the pre-charge voltage on the BL-R bit line has been discharged through transistors **M6** and **M7**. The sense

amplifier receives as inputs a voltage level on output node **33** and BL-R and also receives a reference voltage level  $V_{ref}$ . The sense amplifier **40** provides an output  $D_{out}$ , that is indicative of the relative voltage levels on BL-R and  $V_{ref}$ . If the logic level of the digital information stored by the storage circuitry **22** is logic level 0, then the pre-charge voltage on BL-R will not discharge through transistors M6 and M7. The relative values of  $V_{ref}$  and the BL-R line will cause  $D_{out}$  to have a first sensed value. Conversely, if the logic level of the digital information stored by the storage circuitry **22** is logic level 1, then the pre-charge voltage on BL-R will discharge through transistors M6 and M7. The relative values of  $V_{ref}$  and the BL-R line will cause  $D_{out}$  to have a second sensed value.

The  $V_{ref}$  level is selected based upon a trade-off between read speed and noise margin. During a read cycle, for example, if  $V_{ref}$  is set at  $V_{DD}-V_{tn}$ , where  $V_{tn}$  is NMOS transistor threshold voltage, then only after the voltage level of BL-R drops lower than  $V_{DD}-V_{tn}$ , will the second value of  $D_{out}$  be generated. Alternatively, for example, in order to enhance the read speed,  $V_{ref}$  can be set to  $V_{DD}-0.1V$ . In that alternative case, the read speed would be increased, but the noise margin would be reduced. For example, a MOS diode can be used to select  $V_{ref}$  by producing a diode voltage drop between a  $V_{DD}$  power line and a  $V_{ref}$  line.

FIG. 4 is an illustrative timing diagram used to explain the operation of the circuit of FIG. 3. During time interval T0, the storage circuitry **22** stores a logical 0 value (logic low), and the BL-R bit line including the output node **33** is pre-charged to the supply voltage level  $V_{DD}$ . During time interval T1, a logic 1 value (logic high) is written to the storage circuitry **22**. During time interval T2, the stored logic 1 value is read from the storage circuitry **22**. During time interval T3, a logic 0 value is written to the storage circuitry **22**. During time interval T4, the stored logic 0 value is read from the storage circuitry **22**.

More specifically, during time interval T0, BL-W is 0V; WL-W is 0V; and WL-R, including the output node **33**, is 0V. BL-R is pre-charged to a voltage level  $V_{DD}$ .  $D_{out}$  is 0V. The voltage at storage terminal **28/30** of the storage circuitry **22** is 0V. The voltage  $V_A$  at pull-up node A, the junction of the source terminals of NMOS M1 and PMOS M2, is at a level described below with respect to the description of operation during time interval T3. The voltage  $V_B$  at pull-down node B, the junction of the source terminals of NMOS M3 and PMOS M4, is at 0V.

During time interval T1, a write cycle writes logic level 1 (high) data into the storage circuitry **22**. During the write cycle, the BL-R line can be floating, although it is shown to be at  $V_{DD}$  in the voltage timing diagram of FIG. 4. The WL-R control signal provided to the gate of M6 is at a logic 0 (low) level, turning off M6, thereby de-coupling and electrically isolating the BL-R bit line output, and the output node **33**, from the storage node **28/30**. The BL-W bit line, including input node **31**, provides a logic 1 (high) signal at a voltage level  $V_{DD}$ . The WL-W control signal provided to the gate of M5 is high so as to turn on M5.

In a present embodiment, when M5 is turned on, the input voltage level at input node **31** is communicated to the gate of NMOS transistor M1 and to the gate of PMOS transistor M4. Also, in a present embodiment, when M5 is turned on, the input voltage level at input node **31** is communicated to storage node **28/30**. In one embodiment, M5 is an ordinary enhancement transistor, and the peak voltage of the WL-W control signal is  $HV_{DD}$  which is a voltage boosted one threshold voltage above  $V_{DD}$  so that the voltage level provided at input node **28** can be full  $V_{DD}$ . However, a voltage level of  $V_{DD}$  can be used as the WL-W control signal turn on voltage

level if M5 is implemented as a depletion mode transistor or as a leaky enhancement transistor.

It will be appreciated that even if M5 is an ordinary enhancement transistor, if M1-M4 are depletion transistors, the voltage level of node **28/30** is pulled up to  $V_{DD}$  due to a self-regeneration (self-latch) operation. If M1-M4 are self-regenerative, HVdd is not actually needed. Nevertheless, HVdd ordinarily is desirable in order to transfer to full Vdd from the write bit line (BL-W) to node **28/30**. In one alternative embodiment, transistor M5 is implemented as a depletion mode device or as a leaky enhancement device.

With the provision of a logic 1 value input to the gate of NMOS M1, M1 turns on. Initially, the sources of M1 and M2 interconnected at pull-up node A are at the voltage level  $V_A$ . The voltage at node A rises to  $V_{DD}$  without any voltage drop since M1 is a depletion NMOS transistor. This is also true if M1 is a leaky enhancement transistor. Since the gate of PMOS M2 is coupled to the  $V_{DD}$  supply voltage, and its source is coupled to node A, also at  $V_{DD}$ , the source-to-gate (Vgs) voltage of M2 is 0V. Since PMOS M2 is a depletion PMOS transistor, and Vgs is 0V, there is a conduction path from pull-up node A to the storage node **28/30**. This is also the case when M2 is a leaky enhancement transistor. Therefore, the full supply voltage bias level  $V_{DD}$  is transferred to the storage node **28/30**. Even after the WL-W control signal goes to low (0 V), and turns off transistor M5, the voltage at the storage node **28/30** is maintained at  $V_{DD}$ , maintaining the stored logical 1 level data state.

Also, with the provision of a logic level 1 value to the gate of PMOS transistor M4, M4 is strongly turned off since its Vgs (voltage between its gate and source) is ( $V_{DD}$ ). At the start of time interval T1, the gate of NMOS M3 is coupled to the effective ground voltage bias level  $V_{SS}$  while the pull-down node B voltage is at  $V_{SS}$  (0V in one embodiment). Thus, M3 turns on since the gate-to-source (Vgs) of M3 is  $V_{SS}$  (0V in one embodiment), M4 is strongly turned off (reverse biased), and the leakage current flowing through M4 is smaller than that of M3. Specifically, since M3 is a depletion NMOS transistor with Vgs=0V, there is a conduction path between storage node **28/30** and node B. Consequently, leakage current flows through M3 until a voltage at node B rises to a level of  $V_B < V_{DD}$  at which NMOS M3 is self-reverse biased shutting off the leakage current, whereupon both M3 and M4 both are reversed biased and leakage current is significantly suppressed. When the voltage of node B is  $V_B$ , the Vgs of M4 is ( $V_{DD}-V_B$ ), and the Vgs of M3 is  $-V_B$ . Therefore,  $V_B$  represents a steady-state voltage of node B when the stored data value stored at storage node **28/30** is logic level 1 (high).

The value of  $V_B$  at which M3 becomes self-reverse biased depends on the relative channel conduction strengths of M3 and M4, and can be calculated using Kirchoff's law at pull-down node B. For example, if the channel conductance of M4 was reduced, then the value of  $V_B$  at the point where M3 would become reverse biased also would be increased since more voltage would drop across M4.

Thus, the leakage current from the storage node **28/30** to ground through M3 and M4 can be suppressed to a negligible level. By proper sizing of transistors M3 and M4, node B voltage  $V_B$  can be controlled to suppress the standby leakage current for logic level 1 (high) data storage. More specifically, the voltage  $V_B$  of node B is determined by the ratio of leakage current flowing through M3 and M4. In general, there is an optimum ratio to minimize leakage current. For instance, decreasing the size of M3 relative to the size of M4 decreases the conductance of M3 relative to the conductance of M4 which can lower the level of  $V_B$  at which M3 becomes reverse biased when storing a logic level 1 value by some amount

## 11

$\Delta V_B$ . One result of such decreased relative size of M3 is that during storage of a logic level 1 value, M4 is more highly reversed biased by an amount  $\Delta V_B$  while M3 is less highly reverse biased by the same amount  $\Delta V_B$ . There should be an optimum condition (i.e. an optimum value of  $V_B$ ) at which the degree of reverse biasing of M3 and M4 minimizes leakage current. This optimum level can be set by appropriate relative sizing of M3 and M4.

Therefore, during time interval T1, a logic level 1 value is written into the storage circuitry 22. The logic level 1 (high) input on BL-W turns on the depletion NMOS M1 and turns off the depletion PMOS M4. The voltage  $V_A$  rises to a level at which  $V_{sg}$  of depletion PMOS transistor M2 is 0V, causing M2 to conduct the full supply voltage to the storage node 28/30. The voltage  $V_B$  rises to a level at which depletion NMOS M3 transistor becomes self-reverse biased. With transistors M1 and M2 turned on and transistors M3 and M4 turned off, the voltage of storage node 28/30 is, in effect, 'pulled up' through pull-up node A to the supply voltage bias level. Thus, in a present embodiment, the full supply voltage is provided to maintain the stored logic level 1 signal while leakage current is suppressed through reverse bias of both NMOS M3 and PMOS M4.

During time interval T2, a read cycle reads the stored high logic level (logic 1) signal from the storage circuitry 22. In the example illustrated in FIG. 4, the BL-W bit line, including input node 31, provides a logic level 0 (low) signal, although the signal level on BL-W is unimportant during the read cycle and thus, BL-W can be floating during read operation. The WL-W control signal provided to the gate of M5 is low so as to turn off NMOS transistor M5, thereby electrically isolating the input node 28/30 from the BL-W input. The WL-R control signal is at a logical 1 (high) level causing NMOS transistor M6 to turn on. The voltage level applied to the gate of NMOS transistor M7 is the voltage level representing the data stored by the storage circuitry 22. Specifically, the voltage level of storage node 28/30 is applied to the gate of M7.

In the example of FIG. 4, during time interval T2, the stored data is represented by a logic level 1 (high) voltage, and M7 is turned on due to the stored logic level high data. Throughout the read cycle, the voltage levels  $V_A$  at pull-up node A and  $V_B$  at pull-down node B are maintained at levels consistent with the storage of a logic level 1 (high) at storage node 28/30. During the read cycle, however, the voltage level on the BL-R bit line, including the output node 33, drops from the pre-charge level to a lower voltage level, for example,  $V_{SS}$ .

At the beginning of time interval T2 when the read cycle begins, the BL-R, including output node 33, bit line is at a pre-charge voltage level, which is  $V_{DD}$  in the illustrated embodiment. During the read cycle, the pre-charge voltage on BL-R, including output node 33, discharges to the effective ground level  $V_{SS}$  through M6 and M7. In response to such discharge, the  $D_{out}$  output of the sense amplifier 40 changes from low to high, the second  $D_{out}$  value, indicating a read of a logic level 1 (high) value from storage node 28/30.

More particularly, M6 and M7 cooperate in the reading of stored information from the storage circuitry 22. Transistor M6 functions to determine when a read cycle is to occur. When the WL-R control signal is low, M6 is turned off, and M7 and the storage circuitry 22 are isolated from the BL-R output. When the WL-R control signal is high, as during time interval T2, M6 is turned on, and M7 and the storage circuitry 22 are coupled to the BL-R output. Transistor M7 functions to determine whether or not the pre-charge voltage on BL-R will discharge through M6 and M7 during a read cycle. When M7 is in a turned on state during a read cycle, the pre-charge voltage discharges, and when M7 is in a turned off state

## 12

during a read cycle, the pre-charge voltage does not discharge. The on/off state of M7 is controlled by the voltage level of storage node 28/30 of the storage circuitry 22. If the voltage level maintained by the storage circuitry 22 at storage node 28/30 is low (logic level 0), then M7 is turned off. If the voltage level maintained by the storage circuitry 22 at storage node 28/30 is high (logic level 1) as during time interval T2, then M7 is turned on.

During the read cycle of time interval T2, both M6 and M7 are turned on. The pre-charge voltage on the BL-R bit line output discharges through M6 and M7. The sense amplifier 40 senses the change in the voltage level on the BL-R bit line, including output node 33, and provides a  $D_{out}$  signal having a second value that is indicative of storage of a high logic level signal by the storage circuitry.

In a present embodiment of the invention, the on/off state of M7 is determined by the voltage level maintained by the storage circuitry 22. Transistor M7 in essence serves as an output circuit for the storage circuitry 22 during the read cycle. The on/off state of M7 is indicative of the voltage level maintained by the storage circuitry 22. If M7 is turned on, then a high level is stored. If M7 is turned off, then a low level is stored. Thus, transistor M7 functions to make connection of a voltage discharge path for the BL-R bit line, and output node 33, dependent upon the logic level stored at the storage node 28/30 of the storage circuitry 22.

In general, when M6 is turned on and the storage node voltage is low, transistor M7 communicates the low storage node voltage to output node 33 by preventing discharge of the pre-charge voltage on output node 33. Conversely, when M6 is turned on and the storage node voltage is high, transistor M7 communicates the high storage node voltage by permitting discharge of the pre-charge voltage from output node 33 to effective ground.

The first embodiment of FIG. 3 uses separate read paths and write paths. Specifically, data is written via transistor M5. Data is read via transistors M7 and M8. As explained above, the separate write and read paths promote improved circuit stability. Larger size transistors, with large current carrying capability, can be used to implement M7 and M8 so as to increase read speed. Alternatively, low  $V_t$  transistors can be used to implement M7 and M8 to increase read speed.

During time interval T3, a write cycle writes logic level 0 (low) data into the storage circuitry 22. In the example illustrated in FIG. 4, the BL-R bit line is pre-charged to the  $V_{DD}$  supply bias voltage level, although the voltage level on BL-R is unimportant during the write cycle and thus, BL-R can be floating during the read operation. The WL-R control signal provided to the gate of M6 is at a logic 0 (low) turning off M6, thereby de-coupling and electrically isolating the BL-R bit line, and output node 33, from the storage node 28/30. The BL-W bit line, including input node 31, provides logic 0 (low) signal at a voltage level 0V. The WL-W control signal provided to the gate of M5 is high so as to turn on M5. As explained above with reference to write cycle T1, the peak voltage of the WL-W control signal preferably is  $HV_{DD}$  so that the voltage level communicated to node 28/30 will be full  $V_{DD}$ .

With the provision of a logic 0 value input to the gate of PMOS M4, M4 turns on. The sources of M3 and M4 interconnected at pull-down node B initially are voltage level  $V_B$ , since a high logic level signal previously had been maintained at storage node 28/30. However, since the gate of depletion PMOS transistor M4 receives 0V input, the voltage of pull-down node B drops to the effective ground bias voltage level during time interval T3. This is also true if M4 is a leaky enhancement transistor. Since the gate of NMOS M3 is

coupled to the effective ground bias voltage  $V_{SS}$ , the gate-to-source (Vgs) voltage of M3 is 0V. Since NMOS M3 is a depletion NMOS transistor, and Vgs is 0V, there is a conduction path from the storage node 28/30 to pull-down node B. This is also the case when M3 is a leaky enhancement transistor. Therefore, the storage node 28/30 is coupled to effective ground bias voltage. Even after the WL-W control signal goes to low (0 V), and turns off transistor M5, the voltage at the storage terminal 28/30 is maintained at the effective ground bias voltage  $V_{SS}$ , maintaining the stored logical 0 level data state.

Also, with the provision of a logical 0 value to the gate of NMOS transistor M1, M1 turns off. Specifically, for example, if  $V_{SS}=0V$  is applied to the gate of M1 while the voltage at pull-up node A is at  $V_{DD}$ , then Vgs of M1 is  $-V_{DD}$ , causing M1 to be strongly turned off. Meanwhile, initially the Vgs of M2 is 0V since the gate of M2 is coupled to receive  $V_{DD}$  and the source of M2 is coupled to pull-up node A which initially is at voltage level  $V_{DD}$ . Since M1 is strongly turned off (reversed biased), the leakage current flowing through M1 is smaller than that flowing through M2. More particularly, PMOS transistor M2 turns on since it is a depletion (or alternatively a 'leaky enhancement') transistor with a source-to-gate (Vsg) voltage of 0V. Thus, there is a conduction path, through M2, from pull-up node A to the storage node 28/30. When the voltage of pull-up node A is  $V_A$ , the Vgs of M1 is  $-V_A$  (assuming  $V_{SS}=0V$ ), and the Vgs of M2 becomes  $(V_{DD}-V_A)$ . Leakage current flows through M2 until  $V_A$  achieves a steady-state level at which both transistors M1 and M2 are sufficiently reverse-biased to suppress further leakage current. Therefore  $V_A$  represents a steady-state voltage level at pull-up node A when a logic level 0 (low) value is stored by the storage circuitry 22. Basically, leakage current flows through M2 until a voltage at pull-up node A falls to a level of  $V_A < V_{DD}$  at which PMOS M2 is self-reverse biased shutting off the leakage current, whereupon both M1 and M2 are reversed biased.

The value of  $V_A$  at which M2 becomes self-reverse biased depends on the relative channel conduction strengths of M1 and M2, and can be calculated using Kirchoff's law at node A. For example, if the channel conductance of M2 was reduced relative to the channel conductance of M1, then the value of  $V_A$  at the point where M2 would become reverse biased also would be increased since more voltage would drop across M2.

Thus, the leakage current from the supply bias voltage  $V_{DD}$  to the storage node 28/30 through M1 and M2 can be suppressed to a negligible level. By proper sizing of transistors M1 and M2, pull-up node A voltage  $V_A$  can be controlled to suppress the standby leakage current for logic level 0 (low) data storage. More specifically, the voltage  $V_A$  of pull-up node A is determined by the ratio of leakage current flowing through M1 and M2. In general, there is an optimum ratio to minimize leakage current. For instance, decreasing the size of M1 relative to the size of M2 decreases the conductance of M1 relative to the conductance of M2 which can lower the level of  $V_A$  at which M2 becomes reverse biased when storing a logic level 0 value by some amount  $\Delta V_A$ . One result of such decreased relative size of M1 is that during storage of a logic level 0 value, M2 is more highly reversed biased by an amount  $\Delta V_A$  while M1 is less highly reverse based by the same amount  $\Delta V_A$ . There should be an optimum condition (i.e. an optimum value of  $V_A$ ) at which the degree of reverse biasing of M1 and M2 minimizes leakage current. This optimum level can be set by appropriate relative sizing of M1 and M2.

Therefore, during time interval T3, a logic level 0 value is stored. The logic level 0 (low) input on BL-W, and input node

31, turns on the depletion PMOS transistor M4 and turns off the depletion NMOS transistor M1. The voltage at pull-up node A falls to a voltage level  $V_A$  at which depletion PMOS transistor M2 becomes reverse biased and turns off. The voltage at pull-down node B falls to a level of 0V, and depletion NMOS transistor M3 is turned on. With transistors M1 and M2 turned off and transistors M3 and M4 turned on, the voltage of storage node 23/30 is, in effect, 'pulled down' through pull-down node B to the effective ground voltage bias level. Thus, in a present embodiment, storage node 28/30 is coupled to effective ground bias voltage which serves to maintain the stored logic level 0 signal, while leakage current is suppressed through reverse bias of both NMOS M1 and PMOS M2.

During time interval T4, a read cycle reads the logic level 0 (low) data that is stored by the storage circuitry 22. In the example illustrated in FIG. 4, the BL-W bit line, and input node 31, provide a logic level 0 (low) signal, although the BL-W control signal level on BL-W is unimportant during the read cycle. The WL-W control signal provided to the gate of M5 is low so as to turn off NMOS transistor M5, thereby electrically isolating the storage node 28/30 from the BL-W input. The WL-R control signal is at a logical 1 (high) level causing NMOS transistor M6 to turn on. The voltage level applied to the gate of NMOS transistor M7 is the voltage level representing the data stored by the storage circuitry 22. Specifically, the voltage level of storage node 28/30 is applied to the gate of M7.

In the example of FIG. 4, during time interval T4, the stored data is represented by a logic level 0 (low) voltage, and M7 is turned off due to the stored logic level low data. Throughout the read cycle, the voltage levels  $V_A$  at pull-up node A and  $V_B$  at pull-down node B are maintained at levels consistent with the storage of a logic level 0 (low) at storage node 28/30. Also, during the read cycle, the voltage level on the BL-R bit line, and on output node 33, does not change and remains at the pre-charge voltage level,  $V_{DD}$ .

As explained above, M6 and M7 cooperate in the reading of stored information from the storage circuitry 22. During time the read cycle of time interval T4, transistor M6 is turned on, and transistor M7 is turned off. Therefore, transistor M7 functions to block discharge of the pre-charge voltage  $V_{DD}$  on the BL-R bit line. The sense amplifier senses no change in the voltage level on the BL-R bit line and, and output node 33, provides a  $D_{out}$  signal having a first value that is indicative of storage of a low logic level signal by the storage circuitry 22.

FIG. 5 is an illustrative circuit diagram of a second embodiment of the invention. Components of the seventh embodiment that correspond to identical components of the embodiment of FIG. 3 are identified by primed reference numerals identical to the numerals used to identify corresponding components in FIG. 3. The embodiment of FIGS. 3 and 5 are quite similar, and for that reason, only aspects of the seventh embodiment that are different shall be described.

Basically, the difference between the embodiment of FIG. 3 and that of FIG. 5 is that transistors M5, M6 and M7 in FIG. 5 are depletion type, or alternatively, leaky enhancement type. Whereas, the corresponding transistors of FIG. 3 are ordinary enhancement type devices. An advantage of the use of depletion (or leaky enhancement) transistors as input (M5) and output (M6, M7) devices is avoidance of the use of a boosted voltage  $HV_{DD}$  like that of the embodiment of FIG. 3, reduction of the number of different types of transistors (in this case, all transistors can be implemented with only depletion transistors), and improved read speed.

More particularly, in the embodiment of FIG. 3, in order to transfer the data from BL-W, including input node 31, to the

storage node, **28/30**, with unwanted voltage degradation, a boosted voltage,  $HV_{DD}$ , is employed. However, in the embodiment of FIG. **5**, transistor **M5** is implemented as a depletion (or leaky enhancement) transistor and such boosted voltage is not called for. As a result, a special circuit to generate a boosted voltage is not needed, and unnecessary power consumption for the circuit can be eliminated. One disadvantage of the boosted voltage used in the embodiment of FIG. **3** is that this voltage is maintained in a standby mode, and can be another source of static power consumption. Also, by implementing **M6** and **M7** as depletion transistors, the read speed can be improved due to increased current driving capability for the same input voltage. Basically, since depletion transistors **M6**, **M7** have lower threshold voltages, for the same input voltage, the current driving capability is much larger than that of ordinary enhancement transistors for which the threshold voltage is larger.

One challenge confronted with the use of depletion or leaky enhancement device as an input transistor **M5** is that a depletion device turns on when  $V_{gs}=0.0V$ . Thus, when **M5** is implemented as a depletion (or leaky enhancement transistor), it is desirable to pull the gate voltage of **M5** below  $0.0V$  to ensure turn off of **M5** strongly.

The first S/D of **M7** is coupled to  $V_{ref1}$  which is higher than  $V_{ss}$  by  $V_t$  of enhancement transistor. So, even though stored data is low, **M7** is turned off and BL-bar level is not discharged. But when stored data is high, BL-bar level can be discharged faster due to increased current driving capability of **M6** and **M7**.

FIG. **6A** is an illustrative drawing of a conventional word line driver circuit **50** that can be used with data storage cell circuitry of embodiments of the present invention. FIG. **6B** is an illustrative drawing of a virtual ground signal applied to the driver of FIG. **6A** during active and standby modes of operation. In order to fully turn off transistor **M5** of FIG. **3**, for example, or to suppress leakage current through **M5**, it is desirable to pull down the voltage on a word line (WL) to a lower level in a standby mode than in an active mode. FIG. **6A** shows a driver circuit **50** comprising an inverter **52** with a PMOS transistor **54** and an NMOS transistor **56**. A first S/D of the PMOS device **54** is coupled to a  $V_{DD}$  supply voltage. A second S/D of the PMOS device **54** and a first S/D of the NMOS device **56** are coupled to a data node **58** that is coupled to the WL. A second S/D of the NMOS **56** device is coupled to a virtual ground node **60**. The gates of the PMOS and NMOS transistors **54**, **56** are coupled to an address node **62** which provides address information. In operation, the address information provided to the address node **62** determines the logic level of a signal driven on to WWL by the inverter.

FIG. **6B** shows that a signal  $\phi LV_{SS}$  provided to the virtual ground node **60** is set to  $V_{SS}$  during active mode operation and is set to  $V_{SS}-\Delta V$  during standby mode operation. One advantage to this scheme is that a lower ground voltage can be used during standby mode operation without imparting a speed penalty during active mode operation. One shortcoming with the driver circuit of FIGS. **6A-6B** is that signal  $\phi LV_{SS}$  has limited current driving capability since it is a generated signal rather than a supply such as  $V_{SS}$ , for example. Due to this limited current driving capability, the virtual ground node **60** has a more limited ability to discharge current over a given period of time. During active mode operation, for instance, it is desirable to shut down a word line quickly in order to reduce overall cycle time. The limited driving capability of signal  $\phi LV_{SS}$  can cause delay in discharge of a word line, thereby increasing overall cycle time. In this example, cycle time is the sum of the active cycle time to enable a word line

to perform a given function (e.g., read or write) and the precharge time to disable the word line and set circuits ready for the next operation.

FIG. **7A** is a circuit diagram of an alternative embodiment word line write driver circuit **400**. FIG. **7B** is a signal diagram used to explain the operation of the alternative word line driver **400** both in active and standby modes. The word line driver **400** includes a driver section **402** used to drive the WL control line. The word line driver **400** includes bias circuitry **404** used to regulate a low voltage level provided via WL control line to the gate of write access transistor **M5**.

The driver section **402** includes a PMOS transistor **406** and a first NMOS transistor **408** having respective S/D junctions coupled to form an inverter. An address signal is provided to a driver input node **410**. A WL control signal is provided to WL via a driver output node **412**. One S/D node of PMOS transistor **406** is coupled to a  $V_{DD}$  supply bias voltage source. One S/D node of NMOS transistor **408** is coupled to a virtual ground (LVGND) bias voltage node **414**.

The bias circuitry **404** includes first and second NMOS transistors **M11** and **M12**. Transistors **M11** and **M12** control the voltage level on the LVGND node. One S/D node of **M11** is coupled to the LVGND node, and another S/D node of **M11** is coupled to  $V_{SS}$  supply bias voltage source. A gate of **M11** is coupled to receive a first mode control signal  $\phi LV_{SS}$  provided on a first mode control node **416**. One S/D node of **M12** is coupled to the LVGND node, and another S/D node of **M12** is coupled to receive the first mode control signal  $\phi LV_{SS}$  provided the first mode control node **416**. A gate of **M12** also is coupled to receive a second mode control signal  $\phi STD$  (standby) provided on a second mode control node **418**.

FIG. **7B** shows that in an active mode of operation, first node control signal  $\phi LV_{SS}$  is high, and second node control signal  $\phi STD$  is low. As a result, **M11** is turned on, and **M12** is turned off. The virtual ground bias at the LVGND node is the  $V_{SS}$  voltage level. During active mode operation, when the second NMOS transistor **M11** is turned on and the third NMOS transistor **M12** is turned off, the virtual ground node **414** is coupled the  $V_{SS}$  supply voltage. As a result, there is current discharge capability is enhanced, and performance degradation is reduced. The discharge capability can be further enhanced by ensuring that the second NMOS transistor **M11** is large enough to carry a desired discharge current level and by ensuring that its gate is driven by a voltage somewhat larger than  $V_{DD}$ . FIG. **7B** shows that in a standby mode of operation, first node control signal  $\phi LV_{SS}$  goes below  $V_{ss}$ ,  $V_{SS}-\Delta V$ , and  $\phi STD$  is high. As a result, **M11** is turned off and **M12** is turned on. The virtual ground at the LVGND node is  $V_{SS}-\Delta V$ . The value of  $\Delta V$  is a voltage sufficient to turn off **M5** strongly when **M5** is a depletion transistor.

FIG. **8A** shows another alternative embodiment showing only a bias circuitry portion **500** of a word line write driver circuit that can be used to drive a word line control coupled to the gate of **M5**. One will appreciate that the driver portion (not shown) can be identical to that described with reference to FIG. **7A**. FIG. **8B** is a signal diagram used to explain the operation of the bias circuitry **500**.

The bias circuitry **500** includes NMOS transistors **502** and **504**. One S/D node of transistor **502** is coupled to the LVGND node, and another S/D node of transistor **502** is coupled to  $V_{SS}$  supply voltage. A gate of transistor **502** is coupled to receive a first mode control signal  $\phi_{active}$  provided to a first node control node, i.e., the gate of device **502**. One S/D node of transistor **504** is coupled to a virtual ground node (LVGND), and another S/D node of transistor **504** is coupled to the virtual ground node. A gate of transistor **504** is coupled to

receive a second mode control signal  $\phi_{standby}$  signal provided to a second node control node, i.e., the gate of device **504**.

FIG. **8B** shows that in an active mode, the  $\phi_{active}$  signal is  $HV_{DD}$  ( $V_{DD} + \Delta V_1$ ), and the  $\phi_{LV_{SS}}$  signal is  $V_{SS} - \Delta V_3$ . As a result, in an active node transistor **502** is turned on, transistor **504** is turned off, and the virtual ground node is at  $V_{SS}$ . Thus, there is a discharge path via supply voltage  $V_{SS}$ . Conversely, in a standby mode, the  $\phi_{active}$  signal is  $LV_{SS}$  ( $V_{SS} - \Delta V_2$ ), and the  $\phi_{LV_{SS}}$  signal is  $V_{SS} - \Delta V_1$  when  $\phi_{standby}$  is  $V_{DD}$ . As a result, transistor **502** is turned off, transistor **504** is turned on, and the virtual ground node is at  $V_{SS} - \Delta V_1$ .

The extra-low voltage  $LV_{SS}$  can be produced by on-chip or external negative voltage generator. The generation of the boosted and reduced voltages are well known to persons skilled in the art, form no part of the present invention, and therefore, are not described herein.

Alternatively, for example, **M5** can be implemented as an ordinary enhancement type transistor with a high threshold voltage  $V_t$  which can obviate a need for the circuits of FIGS. **6A**, **7A** or **8A**. In such alternative configuration, **M6** and **M7** can be implemented with low  $V_t$  enhancement transistors to improve the read speed without the requirement of  $V_{ref1}$ .

FIG. **9** is an illustrative circuit diagram of a third embodiment of the invention. Components of the third embodiment that correspond to identical components of the embodiment of FIG. **3** are identified by double primed reference numerals identical to the numerals used to identify corresponding components in FIG. **3**. The embodiment of FIGS. **3** and **9** are quite similar, and for that reason, only aspects of the eighth embodiment that are different shall be described.

The gates of **M2** and **M3** are coupled to receive a reference voltage  $V_{refx}$ . As a result, the gates of **M2** and **M3** are maintained at the same voltage potential. They are at equipotential with respect to each other. A value of  $V_{refx}$  is selected to cause **M2** to turn on more strongly when a high level (logic 1) voltage is stored by the storage circuitry **22**". The value of  $V_{refx}$  is selected so as to also cause **M3** to turn on more strongly when a low level (logic 0) voltage is stored by the storage circuitry **22**". An advantage of turning on **M2** more strongly during storage of a high level voltage and of turning on **M3** more strongly during storage of a low level voltage is improved noise immunity. A data value stored at node **28/30**" can be maintained more safely against possible electrical noise that otherwise might more easily cause discharge or loss of a stored voltage level.

Another purpose of using  $V_{refx}$  is to control the  $V_A$  or  $V_B$ . As explained earlier, by changing the transistor sizes,  $V_A$  and  $V_B$  can be controlled. But in memory cell design, for example, it may be impractical, technically or economically, to change transistor sizes, to adjust  $V_A$  or  $V_B$ , since memory cells typically are repeated many times in an integrated circuit memory. In other words, a seemingly small change in transistor sizes can significantly alter the overall chip. For that reason, it may be more desirable to control  $V_A$  and  $V_B$  by changing a  $V_{refx}$  voltage level rather than by adjusting transistor sizes.

For example, assume that  $V_{refx}$  is selected to be  $\frac{1}{2}V_{DD}$ . During a low-to-high transition such as that illustrated in time interval **T1** of the timing diagram of FIG. **4**, **M1** turns on. The gate voltage of **M2** is  $\frac{1}{2}V_{DD}$ , and the initial source voltage of **M2** is  $V_{DD}$ . Thus, the  $V_{gs}$  of **M2** is  $-\frac{1}{2}V_{DD}$ , and **M2** turns on. Due to the higher magnitude gate voltage ( $V_{refx} = \frac{1}{2}V_{DD}$ ), **M2** of FIG. **9** turns on more strongly during **T1** than does **M2** of FIG. **3** and thereby provides improved noise immunity.

Also, during a low-to-high transition, as in **T1** but with  $V_{refx} = \frac{1}{2}V_{DD}$ , transistors **M3** and **M4** become self reverse biased. In such self reverse biased state, the  $V_{gs}$  of **M3**

becomes  $\frac{1}{2}V_{DD} - V_B$ , and the  $V_{gs}$  of **M4** becomes  $V_{DD} - V_B$ . It will be appreciated that the value of  $V_{refx}$  controls the value of  $V_B$  at which the  $V_{gs}$  of **M3** and the  $V_{gs}$  of **M4** cause both **M3** and **M4** to become reverse biased.

During a high-to-low transition like that of time interval **T3** with  $V_{ref1} = \frac{1}{2}V_{DD}$ , **M4** turns on, and the voltage at pull-down node **B** drops to  $V_{SS}$ . Whereupon, the source voltage of **M3** also drops to  $V_{SS}$ . Assuming that  $V_{SS} = 0V$  the  $V_{gs}$  of **M3** becomes  $\frac{1}{2}V_{DD}$ , and **M3** turns on. Because of the higher magnitude gate voltage ( $V_{refx} = \frac{1}{2}V_{DD}$ ), **M3** of FIG. **9** turns on more strongly during **T3** than does **M3** of FIG. **3** and thereby provides improved noise immunity.

Also, during a high-to-low transition, as in **T3** but with  $V_{refx} = \frac{1}{2}V_{DD}$ , transistors **M1** and **M2** become self reverse biased. In such self reverse biased state, the  $V_{gs}$  of **M1** becomes  $-V_A$  (assuming  $V_{SS} = 0V$ ), and the  $V_{gs}$  of **M2** becomes  $\frac{1}{2}V_{DD} - V_A$ . It will be understood that the value of  $V_{refx}$  controls the value of  $V_A$  at which the  $V_{gs}$  of **M1** and the  $V_{gs}$  of **M2** cause both **M1** and **M2** to become reverse biased.

From the foregoing, persons skilled in the art will appreciate that a value of  $V_{refx}$  can be selected and relative sizing of **M1** and **M2** and of **M3** and **M4** can be provided so as to provide sufficient noise immunity during active state operation while also achieving a sufficient self reverse bias effect to suppress leakage current. Also,  $V_{refx}$  can be used to dynamically control the node voltages  $V_A$  and  $V_B$  during self-reverse bias conditions. Adjusting the value of  $V_{refx}$  can be used to adjust the values of  $V_A$  and  $V_B$  during reverse bias conditions.

FIG. **10** is an illustrative schematic diagram of a multi-state storage circuit in accordance with a fourth embodiment of the invention. Components of the multi-state storage circuit embodiment that correspond to identical components of the embodiment of FIG. **3** are identified by triple primed reference numerals identical to the numerals used to identify corresponding components in FIG. **3**. The embodiment of FIGS. **3** and **10** are quite similar, and for that reason, only aspects of the multi-state storage circuit that are different shall be described.

The multi-state storage circuitry **50** can store data at multiple voltage levels, and each voltage level represents a different logical state. Each voltage level serves as a steady-state level which can be sensed using level sensing circuitry, such as that described below with reference to FIG. **11**. In one embodiment, data can be stored at four different voltage levels, each representing a different logic state. The following table provides an illustrative example of correspondence between data storage voltage levels and logical states.

#### Multi-State Storage Example

Data Storage Voltage Level	Logical State
0.0 V	00
0.6 V	01
1.2 V	10
1.8 V	11

Transistor **M8** enables the storage of multiple different voltage levels, each representing a different logical state. In one embodiment, **M8** comprises an NMOS transistor with a source coupled to input node **28/30**", with a drain coupled to the gate of transistor **M1**, and with a gate coupled to pull-down node **B**. As explained above, the source of **M3** and the source of **M4** are coupled at pull-down node **B**.

Thus, the source of **M8** receives an input voltage level  $V_{in}$  at node **28/30**", and the gate of **M8** receives a pull-down node



B voltage level  $V_B$ . In one embodiment, M8 comprises a depletion device that applies an input voltage  $V_{in}$  to the gate of M1 when  $V_{gs}$  of M8 is greater than or equal to 0.0V. However, alternatively, M8 can be implemented as a leaky enhancement transistor or as an ordinary transistor. Different M8 transistor types lead to the provision of different voltage levels to the gate of M1. When M8 is a depletion transistor, an input voltage without any voltage drop can be applied to the gate of M1 since the M8 threshold voltage is 0V. When M8 is an ordinary transistor, the input voltage dropped voltage by the M8 threshold voltage of the ordinary transistor will be applied to the gate of the M1. Therefore, even though the operation of the circuit is explained based on depletion transistors it is important to note that other types of transistors can be used.

A role of M8 is to control the voltage applied to the gate of M1 as a function of the voltage of node B which is set by the applied (written) voltage to the cell. For example, the voltage of  $V_B$  is different for different written data (voltages). The voltage applied to the gate of M1 is adjustable and dependent upon the most recently input voltage provided to node 28/30'''.

In a present embodiment, essentially, the gate of M1 is coupled to storage node 28/30''' through transistor M8. When M8 is turned on, the gate of M1 is coupled to node 28/30'''. Thus, the voltage applied to the gate of M1 is determined by the voltage of  $V_B$  minus the threshold voltage of M8.

In operation, assuming the above example voltages, any one of the four example voltages can be applied to the gate of M4 as  $V_{in}$ . Each of the four input signal voltage levels corresponds to different digital logic information that can be stored as indicated in the above chart. An input voltage  $V_{in}=0.0V$  turns on M4. An input voltage  $V_{in}=0.6V$  turns off M4. An input voltage  $V_{in}=1.2V$  turns off M4. An input voltage  $V_{in}=1.8V$  turns off M4.

M3 is depletion (or leaky enhancement) transistor that turns on when its  $V_{gs}$  is greater than or equal to 0.0V. In the embodiment of FIG. 10, the gate of M3 is coupled to effective ground  $V_{SS}$ , which is 0.0V. Therefore, M3 is turned on for each of the four example voltage levels.

For  $V_{in}=0.0V$ , M4 is turned on and  $V_B$  becomes 0.0V the current path through M3 and M4 causes  $V_B$  to drop to 0.0V. M8 turns on when  $V_B$  falls to 0.0V since at that point,  $V_B-V_{in}=0.0V$ . Whereupon, the input voltage level  $V_{in}=0.0V$  is applied to the gate of M1. Therefore, the voltage at the source node of M1 is 0.0V

For input voltages of  $V_{in}=0.6V$ , 1.2V or 1.8V, M3 is turned on while M4 is turned off. For each of these input voltages, a leakage current through M3 results in a build up of charge at node B. When the voltage at pull-down node B reaches a level  $V_B$  at which  $V_B-V_{in}$  is greater than or equal to 0.0V, M8 turns on.

For  $V_{in}=0.0V$ , M8 turns on when  $V_B=V_0$ . For  $V_{in}=0.6V$ , M8 turns on when  $V_B=V_1$ . For  $V_{in}=1.2V$ , M8 turns on for  $V_B=V_2$ . For  $V_{in}=1.8V$ , M8 turns on for  $V_B=V_3$ .

Transistor M8 clamps the voltage level stored at node 28/30''' to the input voltage level  $V_{in}$  as follows. When M8 turns on with  $V_{in}=0.0V$ , M1 turns on. M2 also turns on since it has  $V_{DD}$  applied to its gate and  $V_{DD}$  applied to its source giving it a  $V_{gs}$  of 0.0V. The level of node B ( $V_B$ ) is set at  $V_0$  (e.g. 0.0V) according to the bias condition and transistor sizes of M3 and M4. However, if current conduction through M1 and M2 causes  $V_{in}$  to begin to rise above 0.0V, M8 will turn off since its  $V_{gs}=V_B-V_{in}$  no longer will be greater than or equal to 0.0V. Thus, when  $V_{in}=0.0V$ , M8, M1 and M2 cooperate to clamp the voltage at node 28/30''' to 0.0V even after transistor M5 is turned off and the storage circuitry 50 is electrically isolated from the BL-W bit line.

Similarly, when M8 turns on with  $V_{in}=0.6V$ , M1 and M2 turn on. The level of node B ( $V_B$ ) is set at  $V_1$  according to the bias condition and transistor sizes of M3 and M4. However, if current conduction through M1 and M2 causes  $V_{in}$  to begin to rise above the voltage set by this condition, M8 will turn off since its  $V_{gs}$  no longer will be greater than or equal to 0.0V. Thus, when  $V_{in}=0.6V$ , M8, M1 and M2 cooperate to clamp the voltage at node 28/30''' to  $V_1$ .

Likewise, when M8 turns on with  $V_{in}=1.2V$ , M1 and M2 turn on. The level of pull-down node B,  $V_B$ , is set at  $V_2$  according to the bias condition and transistor sizes of M3 and M4. For  $V_{in}=1.2V$ , the newly set voltage level of node B is higher than that set by  $V_{in}=0.6V$  due to reduced channel conductance of M4 for higher applied voltage. However, if current conduction through M1 and M2 causes  $V_{in}$  to begin to rise above  $V_2$ , M8 will turn off since its  $V_{gs}$  no longer will be greater than or equal to 0.0V. Thus, when  $V_{in}=1.2V$ , M8, M1 and M2 cooperate to clamp the voltage at node 28/30''' to  $V_2$ .

Finally, when M8 turns on with  $V_{in}=1.8V$ , M1 and M2 turn on. The level of node B,  $V_B$ , is set at  $V_3$  according to the bias condition and transistor sizes of M3 and M4. For  $V_{in}=1.8V$ , the newly set voltage level of pull-down node B is higher than those set by  $V_{in}=0.6$  and 1.2V due to reduced channel conductance of M4 for higher applied voltage. However, if current conduction through M1 and M2 causes  $V_{in}$  to begin to rise above  $V_3$ , M8 will turn off since its  $V_{gs}$  no longer will be greater than or equal to 0.0V. Thus, when  $V_{in}=1.8V$ , M8, M1 and M2 cooperate to clamp the voltage at node 28/30''' to  $V_3$ .

Therefore, the multi-state storage circuitry 50 of FIG. 10 can maintain more than two voltage levels, and each voltage level can represent a different logical state. The different voltage levels are written into the storage circuitry 50 via BL-W and M5 as described above with reference to FIGS. 3 and 4. Circuit techniques used to produce the different voltage levels for  $V_{in}$  may include the use of different generators and/or internal voltage regulators. These techniques do not form a part of the present invention, and therefore, are not explained herein. Likewise, the different voltage levels are read via M7', M6 and BL-R as described above with reference to FIGS. 3 and 4.

Basically, to clamp the bit line discharge level, M7' can be implemented by a PMOS transistor since bit line voltage can be discharged down only to the stored voltage level plus  $V_{tp}$ , where  $V_{tp}$  is the threshold voltage of PMOS M7'. For example,  $V_1$  level is stored at the storage part, when the bit line voltage is discharged to  $V_1+V_{tp}$ , PMOS transistor, M7' is turned off. Similarly, the bit line can be discharged to  $V_0+V_{tp}$ ,  $V_2+V_{tp}$ ,  $V_3+V_{tp}$  for  $V_0$ ,  $V_2$  and  $V_3$ , respectively. Circuit techniques used to read different voltage levels from the storage circuitry 50 include the use of a sense amplifier with different reference voltages or their combinations.

FIG. 11 is an illustrative drawing of a multi-level sense amplifier 600 that can be used with the embodiment of FIG. 10. The sense amplifier 600 includes first through fourth sense amplifiers 602-1 to 602-4 and first through fourth sense amplifier drivers 604-1 to 604-4. The sense amplifier also includes first through fourth NAND gates 606-1 to 606-4 and first through fourth output drivers 608-1 to 608-4. The various components are interconnected as shown by a network of interconnect lines 610.

Thus, the maximum voltage which node S can have is the gate voltage of M7' (voltage level of node 28/30''') plus the threshold voltage of M7', since M7' is a PMOS transistor. As explained above, the voltage of node C is set by the written voltage,  $V_0$ ,  $V_1$ ,  $V_2$  or  $V_3$ . Therefore, the voltage level where BL-R can discharge in the read operation is  $V_{tp}$  when the voltage level of node 28/30''' is 0V, where  $V_{tp}$  is the threshold

voltage of M7'. Since M7' is turned off when the voltage level of S is below  $V_{tp}$ , so the lowest voltage level where BL-R can discharge is  $V_{tp}$ . When a logic state, V1 is stored at node 28/30''', the minimum voltage which BL-R can discharge is  $V_{in}+V1$ . Similarly, the minimum voltage levels when stored voltages are V2 and V3 are  $V_{tp}+V2$  and  $V_{tp}+V3$ , respectively.

In a preferred embodiment, four sense amplifiers 602-1 to 602-4 are used to sense four different states. Each sense amplifier has a different reference level. Sense amplifier 602-1 has a  $V_{ref1}$  set to have a value between  $V_{tp}$  and  $V_{tp}+V1$ . Sense amplifier 602-2 has a  $V_{ref2}$  set to be between  $V_{tp}+V1$  and  $V_{tp}+V2$ . Sense amplifier 602-3 has a  $V_{ref3}$  set to be between  $V_{tp}+V2$  and  $V_{tp}+V3$ . Sense amplifier 602-4 has a  $V_{ref4}$  set to be larger than  $V_{tp}+V3$ . For a certain case, for example, the third amplifier 602-3 using  $V_{ref3}$  can be omitted. Assume, for example, that the voltage of node 28/30''' is 0.0V, thus the BL-R is discharged to  $V_{tp}$ . Since all reference levels used for the amplifiers are higher than  $V_{tp}$ , all outputs of sense amplifiers (SO1, SO2, SO3 and SO4) are low. Since SO1-bar, SO2-bar, SO3-bar and SO4-bar are inverting signals of the outputs of sense amplifiers, only signal data-00 goes to high. Similarly, Data-01, Data-10 and Data-11 go to high when the stored voltage are V1, V2 and V3, respectively.

In essence, the clamp circuitry operates as a limiting circuit that limits a voltage swing on node 28/30''' as a function of an input voltage level  $V_{in}$  most recently applied to node 28/30'''. More specifically, in one embodiment, the clamp circuitry operates as an adjustable voltage limiting circuit that adjustably limits the voltage level on node 28/30''' based upon the input signal voltage level  $V_{in}$  (i.e., 0.0V, 0.6V, 1.2V, or 1.8V) most recently provided on node 28/30''' via input control circuitry 24'''. It will be appreciated that although only four different example voltage levels have been described for  $V_{in}$ , the multi-state storage circuitry 50 can store more than four different voltage levels representing more than four different logical states.

Various modifications to the preferred embodiments can be made without departing from the spirit and scope of the invention. For example, in another alternative embodiment one or more reference voltages  $V_{refA}$  and  $V_{refB}$  can be provided to the gates of M2 and M3, respectively in the embodiment of FIG. 10. Another alternative, for example, is to provide an additional address transistor in series with M5 in the write path of the embodiments of FIGS. 3, 5, 9 or 10. Since multiple write pass gates (M5s) are connected to a given WL-W at the same time in the memory array, when WL-W is enabled, each BL-W line connected to each cell is charged or discharged according to the cell data regardless of its operation. This can cause unwanted power consumption. Therefore, by adding one more additional address transistors in series with the M5 of each cell, it is possible to use that other transistor to write to selected cells that shares the same WL-W, thereby reducing power consumption. Thus, the foregoing description is not intended to limit the invention which is described in the appended claims.

The invention claimed is:

1. An integrated circuit comprising:

a bias voltage supply node;

a virtual ground node;

a first NMOS transistor including a first S/D electrically coupled to the virtual ground node and including a second S/D electrically coupled to the bias voltage supply node and including a gate electrically coupled to a first mode control node capable of receiving a voltage signal distinct from the virtual ground node; and

a second NMOS transistor including a first S/D electrically coupled to the virtual ground node and including a sec-

ond S/D electrically coupled to the first mode control node and including a gate electrically coupled to a second mode control node.

2. The integrated circuit of claim 1,

wherein the bias supply voltage node is electrically coupled to a power supply node maintained at a voltage level;

wherein the first mode control node is coupled to receive a first mode control signal that turns on the first NMOS transistor in an active mode and that turns off the first NMOS transistor in a standby mode; and

wherein the second mode control node is coupled to receive a second mode control signal that turns on the second NMOS transistor in the standby mode and that turns off the second NMOS transistor in the active mode, wherein the first mode control signal is driven to a voltage value lower than the voltage level of the power supply node and the second mode control node when the driver is in the standby mode, turning the first NMOS transistor off and the second NMOS transistor on.

3. An integrated circuit driver circuit comprising:

a first bias voltage node;

a second bias voltage node;

a first mode control node;

a second mode control node; and

an inverter circuit including a PMOS transistor and a first NMOS transistor,

wherein the inverter includes a data node comprising interconnected first source/drains (S/D) of the PMOS transistor and first NMOS transistor,

wherein the inverter includes an inverter control node electrically coupled to gates of the PMOS transistor and first NMOS transistor,

wherein a second S/D of the PMOS transistor is electrically coupled to the first bias node,

a second NMOS transistor including a first S/D electrically coupled to a second S/D of the first NMOS transistor and including a second S/D electrically coupled to the second bias node and including a gate electrically coupled to the first mode control node, and

a third NMOS transistor including a first S/D electrically coupled to the second S/D of the first NMOS transistor and including a second S/D electrically coupled to the first mode control node and including a gate electrically coupled to the second mode control node.

4. The driver circuit of claim 3,

wherein the first bias node is electrically coupled to a first power supply node maintained at a first voltage level;

wherein the second bias node is electrically coupled to a second power supply node maintained at a second voltage level lower than the first voltage level;

wherein the driver circuit is responsive to receiving at the first mode control node and the second S/D of the third NMOS transistor a voltage higher than the second voltage level, and at the second mode control node a voltage lower than the second voltage level by causing the driver to be in an active mode including that the second NMOS transistor is on and the third NMOS transistor is off,

wherein the driver circuit further is responsive to receiving at both the first mode control node and the second S/D of the third NMOS transistor a voltage lower than the second voltage level, and to receiving at the second mode control signal a voltage higher than the voltage received at the first mode control node by causing the driver to be in a standby mode including that the second NMOS transistor is off and the third NMOS transistor is on.

## 23

5. The driver circuit of claim 3,  
wherein the first control node is coupled to receive an  
address information.
6. The driver circuit of claim 3,  
wherein the first control node is coupled to receive an 5  
address signal and  
wherein the data node is coupled to receive a word line  
signal.
7. The driver circuit of claim 3, further including:  
a virtual ground node; 10  
wherein the second S/D of the first NMOS transistor is  
electrically coupled to the virtual ground node;  
wherein the first S/D of the second NMOS transistor is  
electrically coupled to the virtual ground node; and  
wherein the first S/D of the third NMOS transistor is elec- 15  
trically coupled to the virtual ground node.
8. The driver circuit of claim 2,  
wherein the first bias node is electrically coupled to a VDD  
power supply, and  
the second bias node is electrically coupled to a VSS power 20  
supply.
9. An integrated circuit driver circuit comprising:  
a first bias voltage node electrically coupled to a first power  
supply level;  
a second bias voltage node electrically coupled to a second 25  
power supply level lower than the first power supply  
level;  
a first mode control node;  
a second mode control node;  
an address signal line; 30  
a word line;  
a virtual ground node;  
an inverter circuit including a PMOS transistor with a first  
source/drain (S/D) electrically coupled to the word line  
and a first NMOS transistor with a first S/D coupled to 35  
the word line;

## 24

- wherein gates of the PMOS transistor and first NMOS  
transistor are electrically coupled to the address signal  
line;  
wherein a second S/D of the PMOS transistor is electrically  
coupled to the first bias voltage node;  
wherein a second S/D of the first NMOS transistor is elec-  
trically coupled to the virtual ground node;  
a second NMOS transistor including a first S/D electrically  
coupled to the virtual ground node and including a sec-  
ond S/D coupled to the second bias voltage node and  
including a gate coupled to the first mode control node;  
and  
a third NMOS transistor including a first S/D electrically  
coupled to the virtual ground node and including a sec-  
ond S/D electrically coupled to the first mode control  
node and including a gate electrically coupled to the  
second mode control node,  
wherein the driver circuit is responsive to receiving, at the  
first mode control node and the second S/D of the third  
NMOS transistor, a voltage higher than the second  
power supply level, and at the second mode control node  
a voltage lower than the second power supply level by  
causing the driver to be in an active mode including that  
the second NMOS transistor is on and the third NMOS  
transistor is off,  
wherein the driver circuit further is responsive to receiving,  
at both the first mode control node and the second S/D of  
the third NMOS transistor, a voltage lower than the  
second power supply level, and to receiving at the sec-  
ond mode control signal voltage higher than the voltage  
received at the first mode control node by causing the  
driver to be in a standby mode including that the second  
NMOS transistor is off and the third NMOS transistor is  
on.

\* \* \* \* \*